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Transverse Flux and Flux Reversal Permanent Magnet Generator Systems

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11.1 Introduction

There are certain applications, such as direct-driven wind generators, that have very low speeds (15 to 50 rpm) and microhydrogenerators with speeds in the range of up to 500 rpm and power up to a few megawatts (MW) for which permanent magnet (PM) generators are strong candidates, provided the size and the costs are reasonable.

Even for higher speed applications, but for lower power, today's power electronics allow for acceptable current waveforms up to 1 to 2.5 kHz fundamental frequency f_{in} .

Increasing the number of PM poles $2p_1$ in the PM generator to fulfill the standard condition

$$f_{in} = p_1 \cdot n_n; \quad n_n - \text{speed (rps)} \quad (11.1)$$

thus becomes necessary.

Then, the question arises as to whether the PM synchronous generators (SGs) with distributed windings are the only solution for such applications, when the lowest pole pitch for which such windings can be built is about $\tau_{min} \approx 30$ mm, for three slots/pole. And, even at $\tau = 30$ to 60 mm, is the slot aspect ratio large enough to allow for a high enough electrical loading to provide for high torque density? The apparent answer to this latter question is negative.

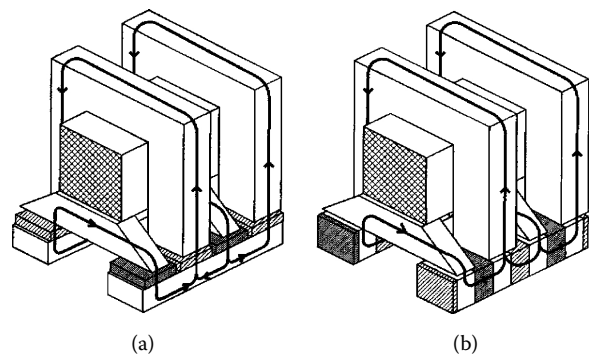


FIGURE 11.1 The single-sided transverse flux permanent magnet (PM) machine: (a) with surface PM pole rotor and (b) with rotor PM flux concentration (interior PM poles).

The nonoverlapping coil winding concept (detailed in [Chapter 10](#)) is the first candidate that comes to mind for pole pitches $\tau > 20$ mm for large torque machines (hundreds of Newton meters [Nm]), but when the pole pitch $\tau \approx 10$ mm, they are again limited in electrical loading per pole, though there is about one slot per pole ($N_s \approx 2p_1$). In an effort to increase the torque density, the concept of a multipole span coil winding can be used, which becomes practical, especially when the number of PM poles $2p_1 > 10$ to 12.

Two main breeds of PM machines were proposed for high numbers of pole applications: transverse flux machines (TFMs) and flux reversal machines (FRMs).

The TFMs are basically single-phase configurations with single circumferential coil per phase in the stator, embraced by U-shaped cores that create a variable reluctance structure with $2p_1$ poles. A $2p_1$ pole surface or an interior pole PM rotor is added (Figure 11.1a and Figure 11.1b). Two or three such configurations placed along the shaft direction would make a two-phase or three-phase machine [1].

There are many other ways to embody the TFM, but the principle is the same. For example, the concept can be extended by placing the PM pole structure in the stator, around the circumferential stator coil, while the rotor is a passive variable reluctance structure with axial or radial airgap (Figure 11.2a and Figure 11.2b) [2].

The PM flux concentration is performed in the stator, in Figure 11.2 configurations, but the PM flux paths in the rotor run both axially and radially; thus, the rotor has to be made of a composite magnetic material (magnetic powder).

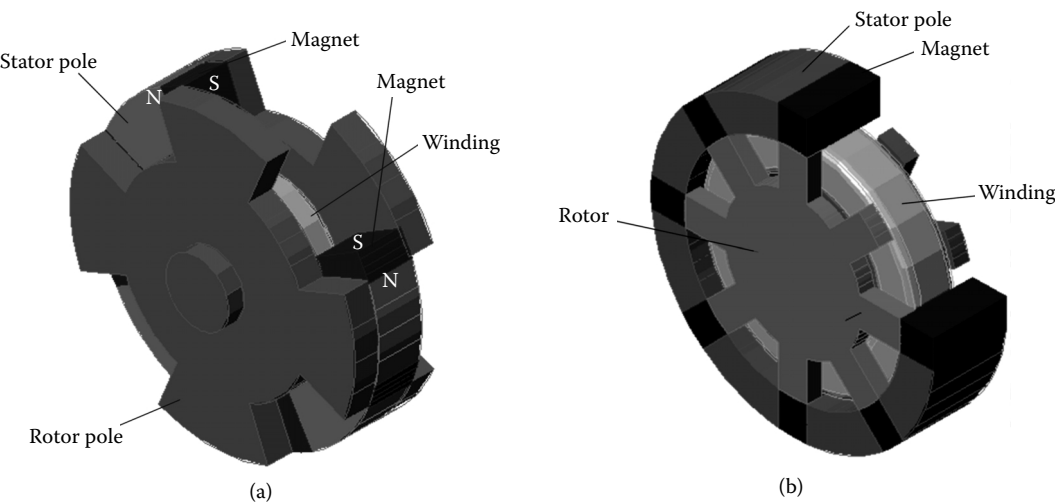


FIGURE 11.2 Transverse flux machine (TFM) with stator permanent magnets (PMs): (a) with axial airgap and (b) with radial airgap.

The TFMs with rotor or stator PM poles are characterized by the fact that the PM fluxes of all North Poles add up at one time in the circumferential coil, and then, after the rotor travels one PM pole angle, all South Poles add up their flux in the coil. Thus, the PM flux linkage in the coil reverses polarity $2p$ times per rotor revolution and produces an electromagnetic field (emf) E_s :

$$E_s = -W_1 \cdot p \cdot \frac{\partial \phi_{PM}^t}{\partial \theta_r} \cdot \frac{d\theta_r}{dt}; \frac{\partial \phi_{PM}^t}{\partial \theta_r} \approx B_g p_1 l_{stack} \cdot \tau \cdot \sin p\theta_r \quad (11.2)$$

where

l_{stack} is the axial length of the U-shape core leg

θ_r is the mechanical angle

B_g is the PM airgap flux density

τ is the pole pitch

Φ_{PM}^t is the total flux per one turn coil

From Equation 11.2,

$$E_s \approx -W_1 \cdot p_1^2 \cdot B_g \cdot \tau \cdot l_{stack} \cdot 2 \cdot \pi \cdot n = -W_1 \cdot B_g \cdot \tau \cdot p_1 \cdot l_{stack} \cdot \omega_1 \cdot \sin p\theta_r \quad (11.2)$$

Equation (11.2) serves to prove that for the same coil and machine diameter ($2p_1\tau = \text{constant}$), the number of turns, and stator core stack length, if the number p_1 of U cores is increased, the emf is increased, for a given speed. This effect may be called torque magnification [2], as torque T_e per phase (coil) is as follows:

$$T_{e_{phase}} = \frac{E_s(\theta_r) \cdot I_1(\theta_r)}{2 \cdot \pi \cdot n}; \quad \omega_1 = 2 \cdot \pi \cdot n \cdot p_1 \quad (11.3)$$

The structure of the magnetic circuit of the TFM is complex, as the PM flux paths are three-dimensional either in the stator or in the rotor or in both.

Soft composite materials may be used for the scope, as their core losses are smaller than those in silicon laminations for frequencies above 600 Hz, but their relative magnetic permeability is below $500 \mu_0$ at 1.0 T. This reduces the magnetic anisotropy effect, which is so important in TFMs. This is why, so far, the external rotor TFM with the U-shape and I-shape stator cores located in an aluminum hub is considered the most manufacturable (Figure 11.3a and Figure 11.3b) [3]. Still, the additional eddy current losses in the aluminum hub carrier are notable.

Though double-sided TFMs with rotor PM flux concentration were proposed (Figure 11.4a and Figure 11.4b) to increase the torque per PM volume ratio, they prove to be difficult to manufacture.

While increasing the torque/volume and decreasing the losses per torque are key design factors, the power factor of the machine is essential, as it defines the kilovoltampere of the converter associated with the TFM for motoring and generating.

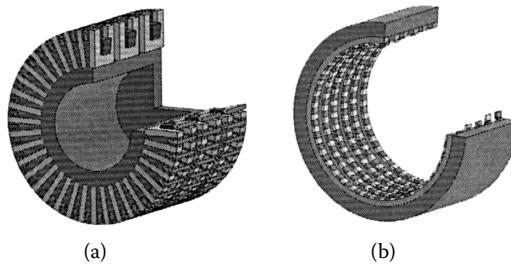


FIGURE 11.3 The three-phase external permanent magnet (PM)-rotor transverse flux machine (TFM): (a) internal stator and (b) external rotor.

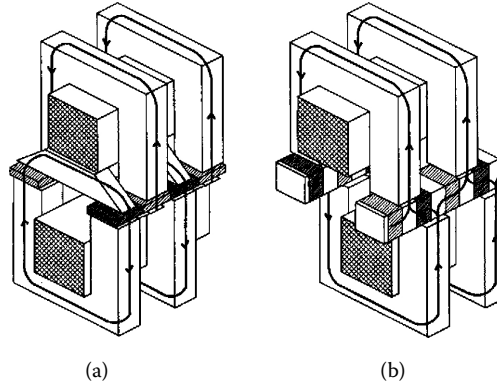


FIGURE 11.4 Double-sided transverse flux machine (TFM) (a) without and (b) with rotor permanent magnet (PM) flux concentration.

The ideal power factor angle ϕ_l of the TFM (or of any PM synchronous machine [SM]) may be defined, in general, as follows:

$$\phi_{ln} = \tan^{-1} \frac{\omega_l \cdot L_s \cdot I_{sn}}{E_s(\omega_l)} \quad (11.4)$$

where

L_s is the synchronous inductance of the machine

I_s is the phase current

Equation 11.4 is valid for a nonsalient pole machine behavior (surface PM pole machines).

For the interior PM pole TFM with flux concentration and a diode rectifier, the machine is forced to operate at about unity power factor. In such conditions, what is the importance of ϕ_{ln} as defined in Equation 11.4? It is to show the power factor angle of the machine for peak torque (pure I_q current control for the nonsalient pole machine). For generality, the presence of a front-end pulse-width modulated (PWM) converter is necessary to extract the maximum power and experience the ϕ_{ln} power factor angle conditions. The power factor at rated current $\cos\phi_{ln}$ is a crucial performance index.

In any case, the larger the machine inductance voltage drop per emf, the lower the power factor goodness of the machine. From this point of view, the PM flux concentration, though it leads to larger torque/volume, also means a higher inductance, inevitably. The claw-pole stator cores were proposed to improve the torque/volume, but the result is still modest [3], due to low power factor. Consequently, only at the same torque density as in the surface PM pole machine, can the TFM with PM flux concentration eventually produce the same power factor at better efficiency and with a better PM usage. While the above rationale in power factor is valid for all PM generators, the problem of manufacturability remains heavy with TFMs.

In order to produce a more manufacturable machine, the three-phase flux reversal PM machine (FRM) was introduced [4]. FRM stems from the single-phase flux-switch generator [5] and is basically a doubly salient stator PM machine. The three-phase FRM uses a standard silicon laminated core with 6k large semiclosed slots that hold 6k nonoverlapping coils for the three phases.

Within each stator coil large pole, there are $2n_p$ PM poles of alternate polarity. Each such PM pole spans τ_{PM} . The large slot opening in the stator spans $2/3 \tau_{PM}$. The rotor has a passive salient-pole laminated core with N_r poles. To produce a symmetric, basically synchronous, machine,

$$2 \cdot N_r \cdot \tau_{PM} = \left(2 \cdot n_p \cdot 6 \cdot k + 6 \cdot k \cdot \frac{2}{3} \right) \cdot \tau_{PM} \quad (11.5)$$

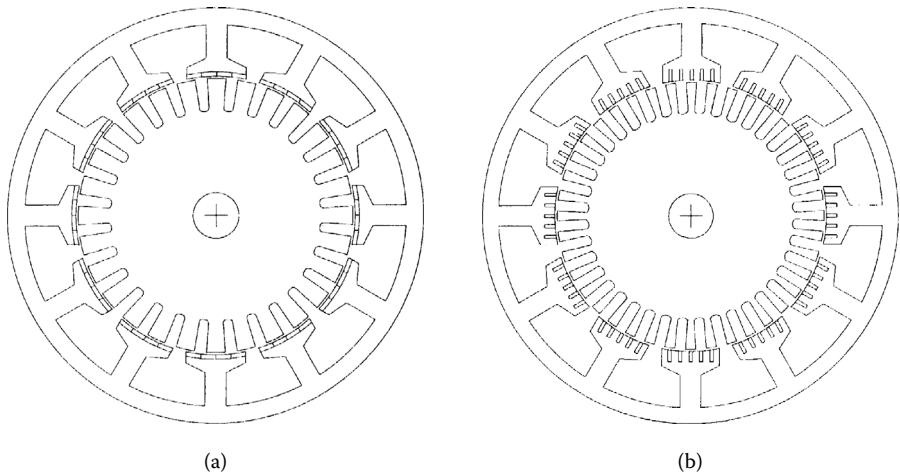


FIGURE 11.5 Three-phase flux reversal machine (FRM) with (a) stator surface permanent magnets (PMs) ($N_s = 12$; $n_p = 2$; $N_r = 28$) and (b) inset PMs ($N_s = 12$; $n_p = 3$; $N_r = 40$).

where
 $k = 1, 2, \dots$
 $n_p = 1, 2, \dots$

A typical three-phase FRM is shown in Figure 11.5a and Figure 11.5b. As for the TFM, the PM flux linkage in the phase coils changes polarity (reverses sign) when the rotor moves along a PM pole span angle. The structure is fully manufacturable, as it “borrows” the magnetic circuit of a switched reluctance machine. However, the problem is, as for the TFM, that the PM flux fringing reduces the ideal PM flux linkage in the coils to around 30 to 60%, in general. The smaller the pole pitch τ_{PM} , the larger the fringing and, thus, the smaller the output. There seems to be an optimum thickness of the PM h_{PM} to pole pitch τ_{PM} ratio, for minimum fringing. On the other hand, the machine inductance L_s tends to be reasonably small, as end connections are reasonable (nonoverlapping coils) and the surface PM poles secure a notably large magnetic airgap. Still, in Reference 4, for a 700 Nm peak torque at 7 N/cm² force density, the power factor would be around 0.3. In order to increase the torque density for reasonable power factor, PM flux concentration may be performed in the stator (Figure 11.6) or in the rotor (Figure 11.7).

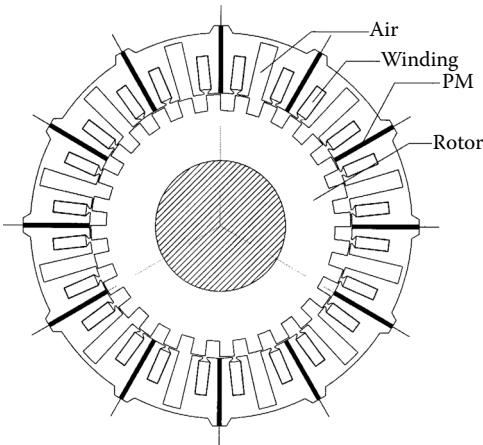


FIGURE 11.6 Three-phase flux reversal machine (FRM) with stator permanent magnet (PM) flux concentration.

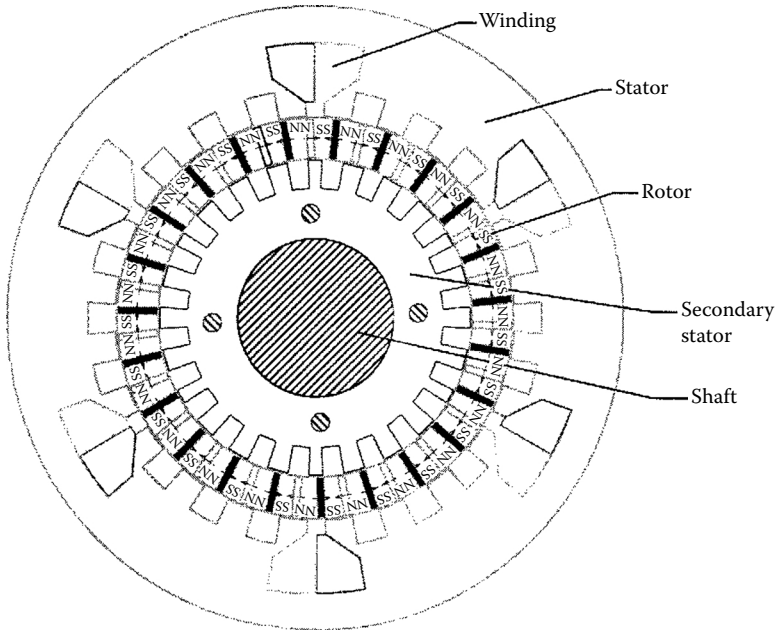


FIGURE 11.7 Three-phase flux reversal machine (FRM) with rotor permanent magnet (PM) flux concentration and dual stator.

The FRM with stator PM flux concentration is highly manufacturable, but as the pole pitch τ_{PM} gets smaller, because the coil slot width w_s is less than τ_{PM} , the power factor tends to be smaller. For $\tau_{PM} = 10$ mm and $w_s = \tau_{PM}$ and a $4w_s$ height, with $j_{peak} = 10$ A/mm² and slot filling factor $k_{fill} = 0.4$, the slot magnetomotive force (mmf) $W_c I_{peak}$ is as follows:

$$W_c \cdot I_{peak} = w_s [mm] \cdot 4 \cdot w_s [mm] \cdot k_{fill} \cdot j_{peak} = 10 \cdot 4 \cdot 10 \cdot 0.4 \cdot 10 = 1600 \text{ Atturns/slot}$$

Larger slot mmfs could be provided for the TFM and FRM with stator surface PM poles. However, the configuration in Figure 11.7 allows for the highest PM flux concentration, which may compensate for the lower $W_c I_{peak}$ and allow for lower-cost PMs, because the radial PM height is generally larger than 3 to 4 τ_{PM} . As with any PM flux concentration scheme, the machine inductance remains large. But, for not so large a number of poles, the machine's easy manufacturability may pay off. On the other hand, the FRM with rotor PM flux concentration configuration (Figure 11.7) provides for large torque density, because, additionally, the allowable peak coil mmf may be notably larger than that for the configurations with stator PM flux concentration (Figure 11.6).

The rotor mechanical rigidity appears, however, to be lower, and the dual stator makes manufacturability a bit more difficult. Still, conventional stamped laminations can be used for both stator and rotor cores.

To provide more generality to the analysis that follows, we will consider only the three-phase TFMs and FRMs, which may work both as generators and motors with standard PWM converters and position-triggered control.

11.2 The Three-Phase Transverse Flux Machine (TFM): Magnetic Circuit Design

The configuration for one phase may be reduced to the one in Figure 11.1a, but inverted, to have an external rotor. The iron behind the PMs on the rotor may be, in principle, solid iron, which is a great advantage when building the rotor.

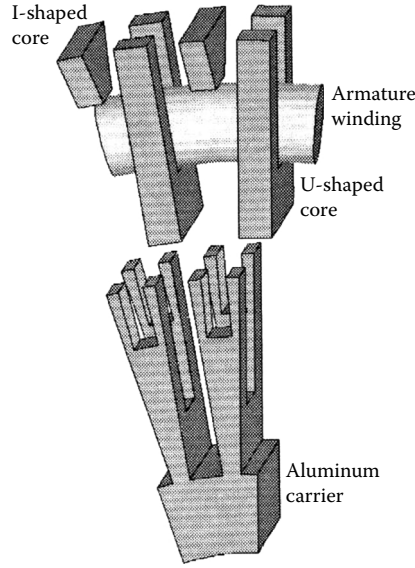


FIGURE 11.8 Transverse flux machine (TFM) — aluminum carrier with interior stator cores and coil.

Also, the stator U-shape and I-cores (Figure 11.4a) may be made of silicon laminations. The aluminum carriers that hold tight the stator I- and U-cores are the main new frame elements that have to be fabricated by precision casting.

Apparently, the circumferential coil has to be wound turn by turn on a machine tool after the U-cores have been implanted in the aluminum carrier. Then, the I-cores are placed one by one in their locations on top of the coil (Figure 11.8).

It was shown that, in order to reduce the cogging torque, the stator U- and I-cores of the three phases have to be shifted by 120 electrical degrees with respect to each other [6]. In such a case, the three phases are magnetically independent, though the PMs are axially aligned on the rotor for all three phases.

As seen in Figure 11.3a, the PM flux paths are basically three-dimensional in the rotor but only bidimensional in the stator. However, as the flux paths are basically radial in the PMs and circumferential-radial in the rotor back iron, the rotor back iron may be made of mild solid steel. The bidimensional flux paths in the stator allow for the use of transformer laminations in the U- and I-cores.

There is, however, substantial PM flux fringing, which crosses the airgap and closes the path between the stator U- and I-cores in the circumferential direction. To reduce it, both U- and I-cores should expand circumferentially less than a PM pole pitch τ_{PM} : $b_u, b_i < \tau_{PM}$. Also, to reduce axial PM flux fringing, the axial distance between the U-core legs l_{slot} should be equal to or larger than the magnetic airgap ($l_{slot} > g + h_{PM}$). But, l_{slot} is, in fact, equal to the width of the open slot where the stator coil is placed.

The U-core yoke h_{yu} and the I-core h_{yi} heights should be about equal to each other and around the value of $2/3 l_{stack}$ in order to secure uniform and mild magnetic saturation in the stator iron cores. Also, the rotor yoke radial thickness h_{yr} should be around half the PM pole pitch, as half of the PM flux goes through the rotor magnetic yoke (Figure 11.9).

We will now approach performance through an analytical method, accounting approximately for magnetic saturation in the stator and iron cores.

Each PM equivalent mmf $F_{PM} = H_c h_{PM}$ (H_c is the coercive field of the PM material) “is responsible” for one magnetic airgap $h_{gM} = g + h_{PM}$. This way, the equivalent magnetic circuit on no load (zero current) is as in Figure 11.9:

$$R_{PM+g} = \frac{h_{PM} + g}{\mu_0 \cdot b_{PM} \cdot l_{stack} \cdot (b_{PM} + b_u)/2} \quad (11.6)$$

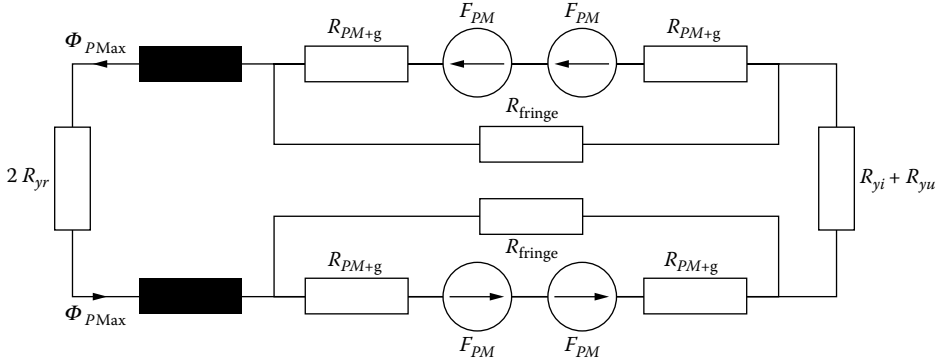


FIGURE 11.9 The magnetic equivalent circuit for permanent magnets (PMs) in the position for maximum flux in the stator coil.

where

R_{PM+g} is the PM and airgap magnetic reluctance

b_{PM} is the PM width ($b_{PM}/\tau_{PM} = 0.66$ to 1.0)

b_u is the U-core width

R_{fringe} is the magnetic reluctance of the PM fringing flux between the stator U- and I-cores through the airgap, corresponding to one leg of the U- and I-cores.

To a first approximation,

$$R_{fringe} \approx \left(\frac{h_{PM}}{\mu_0 \cdot \frac{b_{PM}}{2}} + \frac{\tau_{PM} - b_{yu}}{\mu_0 \cdot \frac{h_{yi}}{2}} \right) \cdot \frac{1}{l_{stack}}; b_{yu} = b_{yi} \quad (11.7)$$

Straight-line magnetic flux paths are considered between U- and I-cores up to the height of the I-core (Figure 11.8). The axial fringing may be added as a reluctance in parallel to R_{fringe} . The stator U- and I-core reluctances R_{yu} and R_{yi} are as follows:

$$R_{yu} = \frac{2 \cdot (h_{slot} + h_{yi})}{\mu_{cu} \cdot b_u \cdot l_{stack}} + \frac{l_{stack} + l_{slot}}{\mu_{yu} \cdot h_{yu} \cdot b_u} \quad (11.8)$$

$$R_{yi} \approx \frac{l_{slot}}{\mu_{yi} \cdot h_{yi} \cdot b_i} + \frac{4 \cdot l_{stack}}{\mu_{yi} \cdot h_{yi} \cdot b_i}; b_i \approx b_u \quad (11.9)$$

Also,

$$R_{yr} = \frac{\tau_{PM}}{\mu_{yr} \cdot b_{yr} \cdot l_{stack}} \quad (11.10)$$

where μ_{cu} , μ_{yu} , μ_{yi} , and μ_{yr} are the magnetic permeabilities dependent on magnetic saturation. As the PM equivalent mmf F_c is given, once all the PM geometry and material properties are known, an iterative procedure is required to solve the magnetic circuit in Figure 11.9. To start with, initial values are given to the four permeabilities in the iron parts: $\mu_{cu}(0)$, $\mu_{yu}(0)$, $\mu_{yi}(0)$, and $\mu_{yr}(0)$. With these values, the flux in the stator and rotor core parts and Φ_{PMax} are computed.

But, the average flux densities in various core parts are straightforward, once Φ_{PMax} is known:

$$\begin{aligned}\phi_{PMax} &= b_{yu} \cdot l_{stack} \cdot B_{cu} = b_{yu} \cdot h_{yu} \cdot B_{yu} = h_{yi} \cdot b_i \cdot B_{yi} \\ \frac{\phi_{PMax}}{2} &= B_{yr} \cdot h_{yr} \cdot l_{stack}\end{aligned}\quad (11.11)$$

Once the average values of flux densities B_{cu} , B_{yu} , B_{yi} , and B_{yr} are computed, and from the magnetization curve of the core materials, new values of permeabilities $\mu_{cu}(1)$, $\mu_{yu}(1)$, $\mu_{yi}(1)$, and $\mu_{yr}(1)$ are calculated. The computation process is reinitiated with renewed permeabilities:

$$\mu(1) = \mu_i(0) + c(\mu_i(1) - \mu_i(0)); \quad c = 0.2 - 0.3 \quad (11.12)$$

such as to speed up convergence.

The computation is ended when the largest relative permeability error between two successive computation cycles is smaller than a given value (say, 0.01). This way, the maximum value of the PM flux per pole (U-core), Φ_{PMax} , in the coil, is obtained.

The PM flux per pole varies from $+\Phi_{PMax}$ to $-\Phi_{PMax}$ when the rotor moves along a PM pole angle ($2\pi/2p_1$ mechanical radians). What is difficult to find out analytically, even with a complicated magnetic circuit model with rotor position permeances, is the variation of Φ_{PM} with rotor position from $+\Phi_{PMax}$ to $-\Phi_{PMax}$ and back.

It was shown by three-dimensional (3D) finite element method (FEM) that the PM flux per pole varies trapezoidally with rotor position (Figure 11.10).

We may consider, approximately, that the electrical angle θ_{const} , along which the PM flux stays constant (Figure 11.10) is rather small and is as follows [3]:

$$\theta_{const} \approx \left(\frac{1}{2} - \frac{(b_{PM} + b_u)}{4 \cdot \tau_{PM}} \right) \cdot \pi \quad [\text{rad}] \quad (11.13)$$

Magnetic saturation alters not only Φ_{PMax} but also the value of θ_{const} , tending to flatten the trapezoidal waveform and bringing it closer to a rectangular waveform of lower height. But, local magnetic saturation plays an even more important role when the machine is under load.

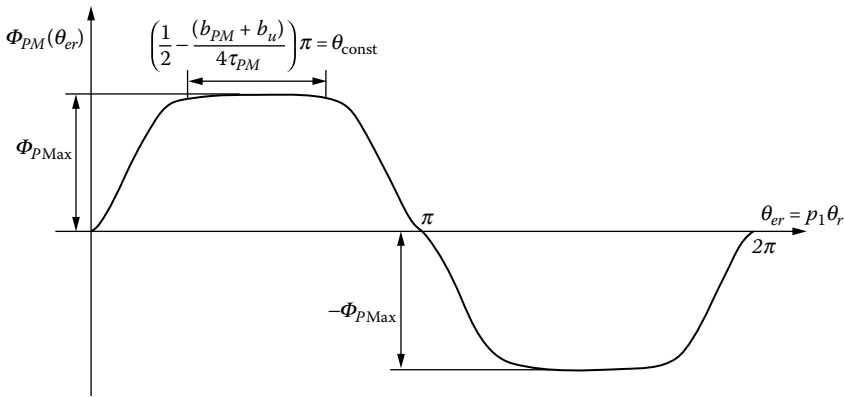


FIGURE 11.10 Permanent magnet (PM) flux per stator pole vs. rotor position.

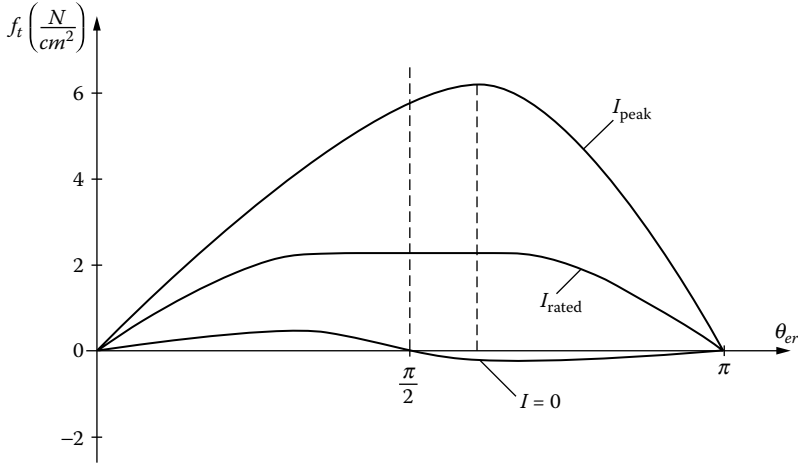


FIGURE 11.11 Typical force density (N/cm²) vs. rotor position and various constant current values.

In general, the computation of the force density (in N/cm²) on the rotor surface is done via the Maxwell stress tensor through FEM. The magnetic saturation presence is evident when the force is calculated for various rotor positions for constant phase current (Figure 11.11).

The PM flux per pole and the emf, obtained through 3D FEM, look as shown in Figure 11.12.

The instantaneous emf per phase E is as follows (Equation 11.2 and Equation 11.3):

$$E = W_1 \cdot p \cdot \frac{d\Phi_{PM}(\theta_{er})}{d\theta_{er}} \cdot 2 \cdot \pi \cdot n \cdot p \quad (11.14)$$

$$\theta_{er} = \omega_r \cdot t \quad (11.15)$$

So, the emf per phase has a waveform in time, which emulates the waveform of $d\Phi_{PM}(\theta_{er})/d\theta_{er}$. The above derivative maximum decreases when the PM pole pitch decreases due to the increase in fringing flux,

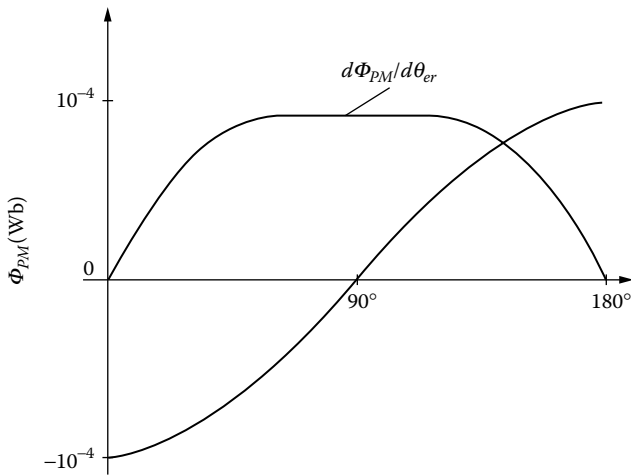


FIGURE 11.12 Typical finite element method (FEM)-extracted permanent magnet (PM) flux and its derivative vs. rotor position.

when the number of poles increases. Consequently, there should be an optimum number of PM poles for a given rotor diameter that produces maximum emf for given W_1 , stack length, and mechanical gap g .

If the phase emf is considered through its fundamental, the average torque per phase, when the emf and current are phase shifted by angle γ_1 , is as follows:

$$(T_e)_{aph} = \frac{E_s \cdot I_1 \cdot \cos \gamma_1}{2 \cdot \pi \cdot n} = \frac{W_1 \cdot p^2 \cdot \phi_{PM_{max}} \cdot I_1 \cdot \cos \gamma_1}{\sqrt{2}} \quad (11.16)$$

where I_1 is the root mean squared (RMS) phase current, and

$$\phi_{PM}(\theta_r) = \phi_{PM_{max}} \cdot \cos(p \cdot \theta_r) \quad (11.17)$$

11.2.1 The Phase Inductance L_s

The phase inductance L_s is composed from the leakage inductance L_{sl} and the main path inductance L_m . As the magnetic circuits of the three phases are separated, there is no coupling inductance between phases. L_s refers to one phase only. The airgap inductance L_m is, approximately,

$$L_m = \mu_0 \cdot \frac{W_1^2}{R_{mag}} \approx \mu_0 \cdot \frac{W_1^2 \cdot p \cdot \frac{b_u + b_{PM}}{2} \cdot l_{stack}}{4 \cdot (g + h_{PM}) \cdot (1 + k_s) \cdot (1 + k_f)} \quad (11.18)$$

where k_s is an equivalent magnetic saturation coefficient that considers the contribution of the iron parts to the total mmf along a flux path. k_f is a fringing coefficient that takes care of the fringing flux in the airgap: $k_f < 0.2$ in general.

In a similar manner, we may treat the maximum flux per pole:

$$\phi_{PM_{max}} = B_{gi} \cdot \frac{b_{PM} + b_u}{2} \cdot l_{stack} \cdot \frac{1}{(1 + k_s)(1 + k_{fringe})} \quad (11.19)$$

k_{fringe} is the PM fringing flux coefficient (very different from k_f) that can go as high as two ($k_{fringe} = 0.7$ to 2), while k_s is the magnetic saturation coefficient. In general, k_s has to be calculated when current is present in stator phases. The leakage inductance may be calculated approximately from slot leakage flux, extended between U-cores, because I-cores “create” a moderate “slot leakage effect”:

$$L_{sl} \approx \mu_0 \cdot W_1^2 \cdot \frac{h_s}{3 \cdot l_{slot}} \cdot p \cdot b_u \cdot (1 + k_i) \quad (11.20)$$

k_i accounts for the leakage inductance between the U-cores. In general, $k_i < 0.2$ to 0.3.

As the total iron area, as seen along the airgap, by the stator coil, is independent of the number of poles, both L_m and L_{sl} are independent of the number of poles $2p_1$. At the same time, the emf E and the torque $(T_e)_{aphase}$ per phase are proportional to the number of poles if the increasing fringing (k_{fringe}) with the number of poles is not considered.

11.2.2 Phase Resistance and Slot Area

The phase resistance R_s is straightforward:

$$R_s \approx \frac{\pi \cdot (D_{is} + 2 \cdot h_{yi} + h_s) \cdot W_1^2}{\sigma_{copper} \cdot \frac{W_1 \cdot l_1}{l_{con}}} \quad (11.21)$$

The ampereturns per phase (RMS value), $W_1 I_1$, may be calculated from Equation 11.16 once the torque per phase requirement is given.

The rated (continuous) current density j_{con} depends on the machine duty cycle, type of cooling, and design optimization criterion (maximum efficiency or minimum machine cost, etc.). In general, $j_{con} = 4$ to 12 A/mm². The window area in the slot A_{slot} is as follows:

$$A_{slot} = h_s \cdot l_{slot} = \frac{W_1 \cdot I_1}{j_{con} \cdot k_{fill}} \quad (11.22)$$

The total slot filling factor k_{fill} is, in general, $k_{fill} = 0.4$ to 0.6. The larger values correspond to the preformed coils eventually made of rectangular cross-sectional conductors.

Example 11.1

Consider sizing a three-phase TFM generator with surface PM interior rotor and single-sided stator (with U- and I-cores) that has the following specifications: $T_{en} = 200$ kNm and $n_n = 30$ rpm.

Solution

While part of the design formulas are included in the previous paragraphs, some new ones are introduced here. They are mainly related to the stator bore diameter D_{is} with given l_{stack}/D_{is} ratio ($\lambda = l_{stack}/D_{is} = 0.05$ to 0.1).

We make use of the force density $f_t = 2$ to 8 N/cm² to determine the interior stator diameter D_{is} :

$$D_{is} = \sqrt[3]{\frac{T_e}{3 \cdot \pi \cdot \lambda \cdot f_t}}; T_e = \frac{D_{is}}{2} \cdot f_t \cdot \pi \cdot D_{is} \cdot 3 \cdot 2 \cdot l_{stack} \quad (11.23)$$

In our case, with $\lambda = 0.05$, $f_t = 2.66$ N/cm²:

$$D_{is} = \sqrt[3]{\frac{200000}{3 \cdot \pi \cdot 0.05 \cdot 2.66 \cdot 10^4}} = 2.525 \text{ m} \quad (11.24)$$

The stack length is as follows:

$$l_{stack} = \lambda D_{is} = 0.05 \cdot 2.515 = 0.1257 \text{ m} \quad (11.25)$$

We now have to choose the airgap g , first.

For such a large diameter, an airgap of 1.5 to 5 mm is required for mechanical rigidity reasons. Let us consider $g = 4 \cdot 10^{-3}$ m. The ideal airgap flux density B_{gi} produced by the PM magnets in the airgap is as follows:

$$B_{gi} = B_r \cdot \frac{h_{PM}}{h_{PM} + g} \quad (11.26)$$

with $B_r = 1.3$ T, $\mu_{rem} = 1.04 \mu_{rem}$ at 100°C (very good NeFeB magnets: $B_{r0} = 1.37$ T at 20°C).

We need to choose a large PM height h_{PM} , because the actual airgap flux density will be notably reduced by the fringing ($k_{fringe} \approx 2$). So, $h_{PM} = g = 3 \cdot 4 \cdot 10^{-3} = 12 \cdot 10^{-3} = 1.2 \cdot 10^{-2}$ m. Consequently, from Equation 11.26,

$$B_{gi} = \frac{1.3 \cdot 12}{12 + 4} = 0.975 \text{ T} \quad (11.27)$$

Now the maximum flux per pole may be calculated if we adopt the number of poles $2p_1$.

In general, the PM pole pitch τ_{PM} should be higher than the total (magnetic) airgap: $g + h_{PM} = 1.6 \cdot 10^{-2}$ m. The number of poles $2p_1$ is, thus,

$$2 \cdot p = \frac{\pi \cdot D_{is}}{\tau_{PM}} = \frac{\pi \cdot 2.515}{0.02} = 394.8 \quad (11.28)$$

Let us consider $2p = 400$ poles and $\tau_{PM} = 1.974 \cdot 10^{-2}$ m. The fundamental frequency at 30 rpm is $f_n = p_1 \cdot n_n = 200 \cdot 30 / 60 = 100$ Hz, which is a reasonable value in terms of core losses. Now, the maximum PM flux per pole from Equation 11.19, with $k_{fringe} = 2.33$; $k_s = 0.1$; $b_{PM}/\tau_{PM} = 0.85$; $b_u/\tau_{PM} = 0.8$ is as follows:

$$\begin{aligned} \phi_{PMax} &= B_{gi} \cdot \frac{b_{PM} + b_u}{2} \cdot l_{stack} \cdot \frac{1}{(1 + k_{fringe}) \cdot (1 + k_s)} \\ &= 0.975 \cdot 1.628 \cdot 10^{-2} \cdot \frac{0.2515}{2} \cdot \frac{1}{(1 + 2.33) \cdot (1 + 0.1)} = 0.5986 \cdot 10^{-3} \text{ Wb} \end{aligned} \quad (11.29)$$

We now turn directly to the torque expression to find the ampere turns per phase $W_1 I_1$ (RMS value) from Equation 11.16, with $\gamma_i = 0$ (emf and current in phase):

$$W_1 \cdot I_1 = \frac{T_{em} \cdot \sqrt{2}}{3 \cdot p^2 \cdot \phi_{PMax}} = \frac{200 \cdot 10^3 \cdot \sqrt{2}}{3 \cdot 200^2 \cdot 0.5986 \cdot 10^{-3}} = 3885 \text{ Aturns} \quad (11.30)$$

This is a reasonable value.

The slot area required for the coil for $j = 6$ A/mm² and $k_{fill} = 0.6$, from Equation 11.22, is as follows:

$$A_{slot} = \frac{W_1 \cdot I_1}{j_{con} \cdot k_{fill}} = \frac{3885}{6 \cdot 0.6} = 1079.16 \text{ mm}^2 \quad (11.31)$$

Taking

$$l_{slot} = 2(g + h_{PM}) = 32 \text{ mm} \quad (11.32)$$

the slot height h_s is

$$h_s = \frac{A_{slot}}{l_{slot}} = \frac{1079}{32} = 33.72 \text{ mm} \quad (11.33)$$

The fact that $h_s/l_{slot} \approx 1$ will lead to a smaller leakage inductance, which is favorable for a good power factor.

Consider $h_{yi} = h_{yu} = 2/3 \cdot l_{stack} = 0.083$ m. Then, from Equation 11.21, the stator phase resistance R_s is

$$R_s = \frac{2.3 \cdot 10^{-8} \cdot \pi \cdot (2.515 + 2 \cdot 0.083 + 0.03372)}{\frac{3885}{6 \cdot 10^6}} \cdot W_1^2 = 0.3026 \cdot 10^{-3} \cdot W_1^2 \quad (11.34)$$

The total winding losses for the machine, P_{con} , are as follows:

$$P_{con} = 3 \cdot R_s \cdot I_1^2 = 3 \cdot 0.3206 \cdot 10^{-3} \cdot 3885^2 = 13.70 \text{ kW} \quad (11.35)$$

The electromagnetic power, P_{elm} , is

$$P_{elm} = T_e \cdot 2 \cdot \pi \cdot n_n = 200 \cdot 10^3 \cdot 2 \cdot \pi \cdot \frac{30}{60} = 628 \text{ kW} \quad (11.36)$$

The winding losses are about 2% of input power.

Now we need to calculate the synchronous inductance $L_s = L_m + L_{sl}$. From Equation 11.19,

$$\begin{aligned} L_m &= \frac{\mu_0 \cdot W_1^2 \cdot p \cdot \frac{b_u + b_{PM}}{2} \cdot l_{stack}}{4 \cdot (g + h_{PM}) \cdot (1 + k_s) \cdot (1 + k_f)} \\ &= \frac{1.256 \cdot 10^{-6} \cdot 200 \cdot 1.62 \cdot 10^{-2} \cdot \frac{0.2515}{2} \cdot W_1^2}{4 \cdot 1.6 \cdot 10^{-2} \cdot (1 + 0.1) \cdot (1 + 0.1)} = \frac{1.321 \cdot 10^{-5}}{2} \cdot W_1^2 [H] \end{aligned} \quad (11.37)$$

Also, from Equation 11.20, L_{se} , the leakage inductance is as follows:

$$\begin{aligned} L_{sl} &= \mu_0 \cdot W_1^2 \cdot \frac{h_s}{3 \cdot l_{slot}} \cdot (1 + k_i) \cdot p_1 \cdot b_u \\ &= 1.256 \cdot 10^{-6} \cdot \frac{0.03372}{3 \cdot 0.032} \cdot (1 + 0.3) \cdot 200 \cdot 0.016 \cdot W_1^2 = 1.834 \cdot 10^{-6} \cdot W_1^2 \end{aligned} \quad (11.38)$$

So,

$$L_{sl} = L_m + L_{se} = \left(\frac{1.321 \cdot 10^{-5}}{2} + 1.834 \cdot 10^{-6} \right) \cdot W_1^2 = 0.844 \cdot 10^{-5} \cdot W_1^2 \quad (11.39)$$

The emf E (rms value) is as follows:

$$E = \frac{T_{em} \cdot 2 \cdot \pi \cdot n}{3 \cdot I_1 \cdot \cos \gamma} = \frac{628 \cdot 10^3}{3 \cdot I_1} = \frac{209.3 \cdot 10^3}{I_1} \quad (11.40)$$

The ideal power factor angle φ_1 (for $\gamma = 0$) is

$$\begin{aligned} \varphi_{1n} &= \tan^{-1} \frac{\omega_1 \cdot L_s \cdot I_1}{E} = \tan^{-1} \left(\frac{2 \cdot \pi \cdot 100 \cdot 0.844 \cdot 10^{-5}}{208.33 \cdot 10^3} \cdot W_1^2 \cdot I_1^2 \right) \\ &= \tan^{-1} (2.544 \cdot 10^{-8} \cdot 3885^2) = \tan^{-1} 0.384 = 21^\circ; \cos \varphi_{1n} = 0.933 \end{aligned} \quad (11.41)$$

The very good value of the power factor angle for E and I in phase is a clear indication of machine volume reduction reserve (the specific tangential force is only 2.66 N/cm²).

The data of the preliminary design are given in [Table 11.1](#).

11.3 TFM — the d - q Model and Steady State

Let us consider here that TFM is provided with a surface PM rotor, and thus, the synchronous inductance L_s is independent of rotor position. We adopt this configuration, as the power factor may be higher due to smaller inductance L_s , even though at slightly smaller torque density than for the PM flux concentration configurations. Though the emf waveform is not quite sinusoidal, we consider it here as sinusoidal.

TABLE 11.1 Example Design Summary

Electromagnetic torque	$T_e = 200 \text{ kNm}$
Speed	$n_n = 30 \text{ rpm}$
Force density	$f_i = 2.66 \text{ N/cm}^2$
Current density	$j_{con} = 9 \text{ A/mm}^2$
Stator interior diameter	$D_{is} = 2.515 \text{ m}$
Stator stack U-core leg	$l_{stack} = 0.1257 \text{ m}$
Mechanical gap	$g = 4 \cdot 10^{-3} \text{ m}$
Permanent magnet radial thickness	$h_{PM} = 12 \cdot 10^{-3} \text{ m}$
Number of permanent magnet poles	$2 \cdot p = 400$
Fundamental frequency	$f_n = 100 \text{ Hz}$
Permanent magnet pole pitch	$\tau_{PM} = 1.974 \cdot 10^{-2} \text{ m}$
Axial room between U-core legs	$l_{slot} = 32 \cdot 10^{-3} \text{ m}$
U-core width	$b_{st}/\tau_{PM} = 0.8$
Permanent magnet width per pole	$b_{PM}/\tau_{PM} = 0.85$
U-core yoke height	$h_{yu} = 2/3 \cdot l_{stack}$
I-core yoke height	$h_{yi} = h_{yu}$
Slot useful height	$h_s = 33.72 \cdot 10^{-3} \text{ m}$
Total U-core interior height	$h_s + h_{yi}$
External stator diameter	$D_{os} = 2.621 \text{ m}$
Total axial length per phase	$l_{slot} + 2 \cdot l_{stack} = 0.032 + 0.2515 \text{ m} = 0.2835 \text{ m}$
Total axial length for three phases	$l_{axial} = 3 \cdot (l_{slot} + 2 \cdot l_{stack}) = 0.85 \text{ m}$
Reduction of permanent magnet flux due to fringing	$1/(1 + k_{fringe}) = 1/3$
Stator resistance per phase	$R_s = 0.454 \cdot 10^{-3} \cdot W_l^2$
Stator inductance per phase	$L_s = 0.844 \cdot 10^{-5} \cdot W_l^2$
Emf at $f_i = 100 \text{ Hz}$	$E = 53.61 \cdot W_l \text{ (rms)}$
Stator copper losses	$P_{con} = 20.55 \text{ kW}$
Electromagnetic power	$T_e \cdot 2 \cdot \pi \cdot n_n = 628 \text{ kW}$
Power factor angle (with E and I in phase)	$\phi_{in} \approx 21^\circ; \cos \phi_n = 0.933$

The d - q model is thus straightforward, if the core losses are neglected for the time being:

$$\begin{aligned}
 \bar{I}_s \cdot R_s + \bar{V}_s &= -\frac{d\bar{\psi}_s}{dt} - j \cdot \omega_r \cdot \bar{\psi}_s \\
 \bar{\psi}_s &= \psi_d + j \cdot \psi_q; \psi_d = \psi_{PM} + L_s \cdot I_d; \psi_q = L_s \cdot I_q \\
 \bar{I}_s &= I_d + j \cdot I_q; \bar{V}_s = V_d + j \cdot V_q; \psi_{PM} = \phi_{PM\max} \cdot W_l \cdot p_1
 \end{aligned} \tag{11.42}$$

where

W_l is the turns per coil (phase)

p_1 is the pole pairs:

$$T_e = \frac{3}{2} \cdot p_1 \cdot (\psi_d \cdot I_q - \psi_q \cdot I_d) = \frac{3}{2} \cdot p_1 \cdot \psi_{PM} \cdot I_q$$

For steady state $\frac{d}{dt} = 0$, and the vector diagram for generating is as shown in Figure 11.13a and Figure 11.13b for pure I_q control ($I_d = 0$) and for unity power factor (with diode rectifier). Pure I_q control for $L_s = \text{constant}$ corresponds to maximum torque per current, but the machine requires reactive power for magnetization. A controlled front rectifier is required.

The performance is acceptable, but the force density is not large enough.

For the diode rectifier, the current always contains an I_d component (Figure 11.13b), and thus, the torque is produced with more losses. However, the reactive power required is zero, and the diode rectifier is less expensive. If constant direct current (DC) link voltage operation at variable speed is required, either an active front or PWM converter is provided on the machine side or a diode rectifier plus a

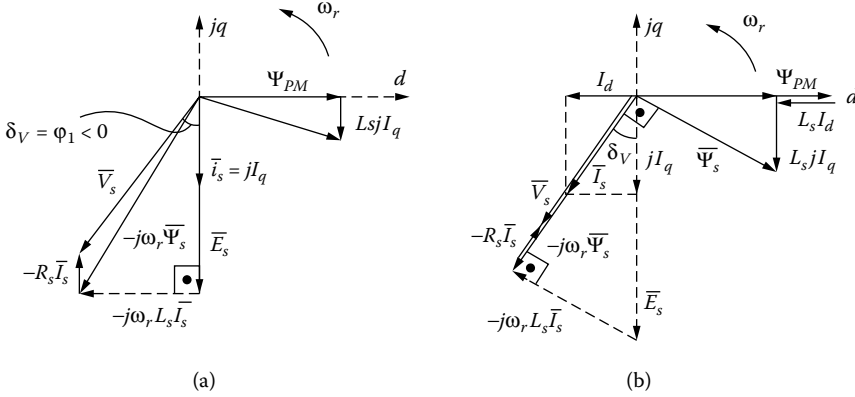


FIGURE 11.13 The transverse flux machine (TFM) vector diagram for steady state: (a) pure I_q control and (b) with power factor control (with diode rectifier).

DC–DC boost converter is required. The core losses might be considered, mainly for steady state, as follows:

$$p_{iron} = \frac{3}{2} \cdot \frac{\omega_r \cdot \Psi_s^2}{R_{iron}} \quad (11.43)$$

In Equation 11.43, the core losses are considered to be produced by the resultant flux in the machine, in the presence of stator current. The core losses p_{iron} might be determined from tests that segregate the iron losses, or they may be calculated from analytical or FEM models, as shown in [Chapter 10](#).

The efficiency may be defined as follows:

$$\eta = \frac{P_{2el}}{P_{2el} + p_{mec} + p_{iron} + p_{copper} + p_{rotor} + p_{Al}} \quad (11.44)$$

The rotor losses p_{rotor} refer to the eddy current losses in the PMs, which may be neglected only for $f_i \leq 100$ Hz, in general. The strayload losses are included in p_{iron} and p_{copper} , but the eddy current losses in the aluminum carriers of the U- and I-cores are individualized as p_{Al} and may be calculated only through 3D FEM eddy current models [7].

When the fundamental frequency goes up, above 100 Hz, p_{iron} and p_{rotor} become important and have to be dealt with using great care (see Chapter 10 for more details).

Example 11.2

Consider the TFM design in Example 11.1, and calculate the number of turns per phase for a line voltage $V_{Ln} = 584$ V (RMS) — star connection — at rated speed for E, I in phase, and for the same number of turns W_l and unity power factor, and calculate the performance for the same current as above.

Solution

We simply make use of the vector diagram (Figure 11.13a), where Ψ_{PM} is

$$\begin{aligned} \Psi_{PM} &= \frac{E(rms) \cdot \sqrt{2}}{\omega_r} = \frac{53.61 \cdot W_l \cdot \sqrt{2}}{2 \cdot \pi \cdot 100} = 0.1203 \cdot W_l \\ V_{sn} &= \frac{V_{Ln}(rms) \cdot \sqrt{2}}{\sqrt{3}} = \frac{584 \cdot \sqrt{2}}{\sqrt{3}} = 476 \text{ V} \end{aligned} \quad (11.45)$$

From Figure 11.13a,

$$E_s = E(rms) \cdot \sqrt{2} = V_s \cdot \cos \phi_1 + R_s \cdot I_s \quad (11.46)$$

or

$$75.59 \cdot W_1 = 476 \cdot 0.933 + 0.454 \cdot 10^{-3} \cdot W_1^2 \cdot I_1 \cdot \sqrt{2} \quad (11.47)$$

With $W_1 I_1 = 3885$ Atturns/coil, we obtain $W_1 \approx 6$ turns/coil. As the core and mechanical losses have not been considered, the efficiency will reflect only the presence of the already calculated copper losses:

$$\eta = \frac{P_{elm} - p_{con}}{P_{elm}} = \frac{P_2}{P_{elm}} = \frac{628 - 13.70}{628} = \frac{614.3 \text{ kW}}{628 \text{ kW}} = 0.978 \quad (11.48)$$

Alternatively,

$$\eta_{en} = \frac{\sqrt{3} \cdot V_{Ln} \cdot I_1 \cdot \cos \phi_1}{P_{elm}} \quad (11.49)$$

with

$$I_1 = \frac{W_1 \cdot I_1}{W_1} = \frac{3885}{6} = 647.5 \text{ A} \quad (11.50)$$

$$\eta_{en} = \frac{\sqrt{3} \cdot 584 \cdot 647 \cdot 0.933}{628 \cdot 10^3} = \frac{609}{628} = 0.971 \quad (11.51)$$

The difference in efficiency from the two formulas above is only due to calculation errors, as the model is the same.

Now, for the second case, we have to notice that if we consider the same phase current (RMS) I_1 and voltage V_{Ln} , the vector diagram in Figure 11.13b provides the following:

$$(V_s + R_s \cdot I_1)^2 + \omega_1^2 \cdot L_s^2 \cdot I_1^2 = E_s^2 \quad (11.52)$$

with $W_1 = 6$ turns, $I_1 = 647.5$ A, we can calculate, from Equation 11.52, the terminal voltage:

$$\begin{aligned} V_s' &= \sqrt{E_s^2 - \omega_1^2 \cdot L_s^2 \cdot I_s^2} - R_s \cdot I_s \\ &= \sqrt{(75.59 \cdot 6)^2 - (2 \cdot \pi \cdot 100 \cdot 1.5044 \cdot 10^{-5})^2 \cdot 6^4 \cdot 647.5^2} \\ &\quad - 0.3026 \cdot 10^{-3} \cdot W_1^2 \cdot I_1 \cdot \sqrt{2} = 396.48 - 9.945 = 386.63 \text{ V} \end{aligned} \quad (11.53)$$

$$V_L' = \frac{V_s' \cdot \sqrt{3}}{\sqrt{2}} = 386.63 \cdot \sqrt{\frac{3}{2}} = 473.40 \text{ V} < 584 \text{ V} \quad (11.54)$$

The line terminal voltage for the same current decreased notably for unity power factor.

Now, the output power is only

$$(P_2)_{\cos \phi_1=1} = \sqrt{3} \cdot 473.40 \cdot 647.5 = 530.29 \text{ kW} \quad (11.55)$$

So, for the same number of turns $W_l = 6$ and the same conductor cross-section and current and speed, with a diode rectifier, the generator will produce 20% less power. The additional reactive power provided through the active front PWM converter, when E and I are in phase, allows for delivering power in better conditions (higher voltage and power).

Note on the Control of TFM

Once a three-phase machine is considered, the control may be implemented with sinusoidal current (for almost sinusoidal emf) or with trapezoidal current (for trapezoidal emf), with third harmonic tapping, if a diode rectifier is used. The control system, with or without motion sensors, is practically the same as that for standard PMSGs, as detailed in [Chapter 10](#). This is why we do not elaborate here on TFM control.

The methods used to reduce the cogging torque are similar to those for standard PMSGs. It is to be seen if the TFM will make it to the markets soon, at least as a low-speed high-torque direct-driven generator.

11.4 The Three-Phase Flux Reversal Permanent Magnet Generator: Magnetic and Electric Circuit Design

As mentioned in the section on TFMs, the large PM flux fringing and, consequently, the low power factor are the main obstacles in the way for producing very good performance, even at very low speeds. The FRMs face similar problems, but they are easier to manufacture [8–11].

We already introduced the PM flux concentration schemes for FRM with stator ([Figure 11.6](#)) and rotor PMs ([Figure 11.7](#)). We will, however, treat in detail here only the surface stator PM configuration, as it is similar to the TFM investigated in the previous paragraph and is characterized by a smaller inductance (in P.U.), though PM flux fringing is large. We will first present a conceptual design and complete FEM investigations of a 200 Nm, 128 rpm (60 Hz) FRM. Then, through a numerical example, a FRM preliminary design methodology is presented for 200 kNm, 30 rpm, 100 Hz system, as in [Example 11.1](#) and [Example 11.2](#). We will refer to very low speeds, as we feel that for speeds above 500 to 600 rpm, PMSGs with distributed or fractional (with tooth-wound coils) windings are already well established ([Figure 11.14](#)).

The low-speed operation at a frequency of 50 to 60 Hz (to secure moderate core loss but still make good use of iron) implies a large number of “poles” or “pole pairs” or “periods” of the flux reversal along the machine periphery. In the FRM with N_r rotor salient poles, the speed n and frequency f_1 are related by the following:

$$f_1 = n \cdot N_r \quad (11.56)$$

The two-pole pitch angle corresponds to two PM poles of alternate polarities on the stator, that is,

$$2 \cdot \tau_{PM} = 2 \cdot \tau_{rot} = \frac{\pi \cdot D_r}{N_r} \quad (11.57)$$

where D_r is the rotor diameter. A stator pole may accommodate $2 \cdot n_p$ PMs. The electrical angle of the space between two neighboring stator pole coils should be 120 electrical degrees or $120^\circ/N_r$, geometrical degrees or $2 \cdot \tau_{PM}/3$. Thus,

$$N_s \cdot \left(2 \cdot n_p \cdot \tau_{PM} + \frac{2 \cdot \tau_{PM}}{3} \right) = \pi \cdot (D_r + 2 \cdot g) \approx 2 \cdot \tau_{PM} \cdot N_r \quad (11.58)$$

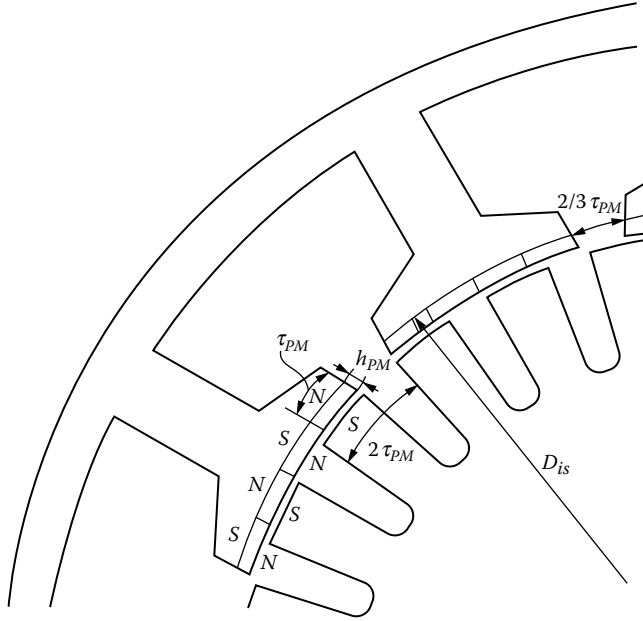


FIGURE 11.14 Surface permanent magnet (PM) flux reversal machine (FRM) geometry details ($N_s = 12$, $n_p = 2$, $N_r = 28$).

where N_s is the number of stator poles, and g is the mechanical airgap. Eliminating τ_{PM} from Equation 11.58 yields the following:

$$N_s \cdot \left(n_p + \frac{1}{3} \right) = N_r \quad (11.59)$$

The typical configuration for $N_s = 12$ is shown in Figure 11.4 and Figure 11.15. The PMs on the stator pole shoes are very close to the airgap. We may say it has pole PMs.

The FRM for low-speed drives has the following distinct features:

- It uses conventional stamped laminations both on the stator and on the rotor.
- It has no PMs or windings on the rotor.
- It has PMs on the stator, where their temperature can be easily monitored and controlled.
- The stator has concentrated coils that are easy to manufacture.
- The lowest pole pitch τ_{PM} is to be larger than (PM + airgap) thickness [8–11] to limit flux fringing in PM utilization, as is the case with the transverse PM flux machine that has the PMs placed on the rotor.
- The higher the rotor diameter (or torque), the larger the maximum number of pole pairs ($2 \cdot \tau_{PM}$), and thus, the lower the speed at 50 (60) Hz.
- The pole PM FRM has lower inductances than the inset PM FRM (Figure 11.15), as expected. Also, the cogging torque of the latter is intrinsically smaller.
- For the inset PM FRM, the airgap should be as small as mechanically feasible, and the PM thickness $h_{PM} > 6 \cdot g$. Also, the stator and rotor teeth should be equal, while the rotor interpole will span the rest of the double-pitch τ_{PM} :

$$\tau_{PM} = b_{t1} + h_{PM}; \quad 2 \cdot \tau_{PM} = b_{t2} + b_{s2}; \quad b_{t2} = b_{t1} \quad (11.60)$$

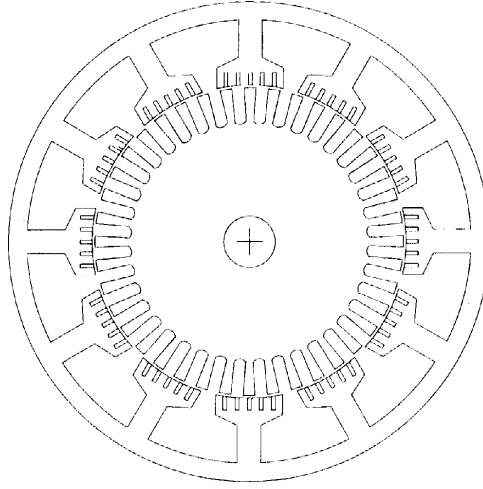


FIGURE 11.15 Low-speed flux reversal machine (FRM) with inset permanent magnets (PMs) on stator.

Moreover, $b_{rl}/h_{PM} > 2/3$ provides enough saliency. Thus, the inset PM FRM with an airgap of 0.2 mm, $h_{PM} = 1.5$ mm, would allow for $b_{rlmin} = 5$ mm or a pole pitch $\tau_{PM} = 5 + 1.5 = 6.5$ mm, which is 65% of the one considered for the pole PM configuration. Further reductions of the airgap and magnet thickness may lead to even smaller pole pitches. Thus, at 60 Hz, the speed is accordingly reduced to a value depending on the rotor diameter.

- The inset PM FRM has the PMs parallel to the stator magnet flux lines and is much more difficult to demagnetize. As a bonus, the flux in the magnets varies less (especially under load). Thus, the eddy currents induced with PMs are notably smaller than those for the pole PM configuration, that has PMs that directly experience the stator current additional field.
- The much lower total (magnetic) airgap of the inset PM FRM leads to much larger inductance, which is limited by heavy saturation of the core for high currents (overload). Therefore, the stator current limits are mainly governed by stator temperature and magnetic oversaturation in the inset PM configuration, and stator temperature and PM demagnetization limit the currents for the pole PM FRM.

11.4.1 Preliminary Geometry for 200 Nm at 128 rpm via Conceptual Design

Consider the pole PM configuration only, with $N_r = 28$ rotor poles, $N_s = 12$ stator poles, with $2n_p = 4$ PMs per stator pole. The designated primary frequency is $f_1 = 60$ Hz. It corresponds to the rated speed $n_n = (f_1/N_r) \cdot 60 = 128.5$ rpm. We start the preliminary design from the specific tangential force f_t to calculate the torque T_e :

$$T_e = f_t \cdot \pi \cdot D_r \cdot l_{stack} \cdot \frac{D_r}{2}; f_m = 1.5 - 4 \text{ N/cm}^2 \quad (11.61)$$

where D_r is the rotor diameter, and l_{stack} is the stack length. A stack-length-to-rotor-diameter D_r ratio λ is defined as follows:

$$\lambda = \frac{l_{stack}}{D_r} = 0.2 - 1.5 \quad (11.62)$$

The lower values of λ are justified for a large number of stator poles. With $N_s = 12$, $\lambda = 1.055$, $T_e = 200$ Nm, and $f_{in} = 2$ N/cm² (continuous duty), from Equation 11.62, we obtain

$$D_r = \sqrt[3]{\frac{2 \cdot T_{en}}{\pi \cdot f_{in} \cdot \lambda}} = \sqrt[3]{\frac{2 \cdot 200}{\pi \cdot 2.0 \cdot 10^4 \cdot 1.055}} = 0.182 \text{ m} \quad (11.63)$$

We may choose $D_r = 0.180$ m to obtain $l_{stack} = 1.055 \cdot 0.18 = 0.19$ m. The PM flux per stator pole Φ_{PM} is as follows:

$$\phi_{PM} = \frac{l_{stack} \cdot B_{PMi} \cdot \tau_{PM} \cdot n_{pp}}{1 + k_{fringe}} \quad (11.64)$$

B_{PMi} is the ideal flux density in the airgap as given by

$$B_{PMi} = B_r \cdot \frac{h_{PM}}{h_{PM} + g} \quad (11.65)$$

where $h_{PM} = 2.5$ mm, and $g = 0.5$ mm. At 75°C, $B_r = 1.21$ T, $H_c = 651 \cdot 10^3$ A/m for neodymium–iron–boron (NdFeB) PMs. k_{fringe} is a flux fringing coefficient to be determined by FEM. We choose here an initial safe value $k_{fringe} = 2.33$. The flux is to vary almost sinusoidally with the rotor position θ_r :

$$\phi_{PM}(\theta_r) = \phi_{PM} \cdot \sin(N_r \cdot \theta_r) \quad (11.66)$$

Therefore, the PM flux per stator pole derivative with θ_r is

$$\frac{d\phi_{PM}(\theta_r)}{d\theta_r} = \frac{l_{stack} \cdot B_{PMi} \cdot \tau_{PM} \cdot n_{pp} \cdot N_r \cdot \cos(N_r \cdot \theta_r)}{1 + k_{fringe}} \quad (11.67)$$

There are $N_s/3$ stator poles per phase and, with all coils per phase in series, according to Equation 11.59, the emf amplitude per phase E_m is

$$E_m = \frac{N_s}{3} \cdot n_c \cdot \frac{2 \cdot \pi \cdot n \cdot l_{stack} \cdot \pi \cdot D_r \cdot n_{pp} \cdot B_{PMi}}{1 + k_{fringe}} \quad (11.68)$$

where n_c is the number of turns per coil. For a given sinusoidal current I (RMS), in phase with the emf (I_q current control), the torque T_e is, again,

$$T_e = \frac{3}{2} \cdot E_m \cdot I \cdot \frac{\sqrt{2}}{2 \cdot \pi \cdot n} \quad (11.69)$$

In our case, we find $n_c I_n \sqrt{2}$:

$$\begin{aligned} n_c \cdot I_n \cdot \sqrt{2} &= n_c \cdot \frac{2}{3} \cdot T_e \cdot \frac{2 \cdot \pi \cdot n}{E_m} = \frac{2}{3} \cdot 200 \cdot \frac{1}{4 \cdot 0.19 \cdot \pi} \cdot \frac{1}{0.18 \cdot 0.3 \cdot 1.21} \\ &\approx 0.855 \cdot 10^3 \text{ Aturns/coil.} \end{aligned} \quad (11.70)$$

For E_m , we used Equation 11.68.

Choosing a rated current density $j_{cn} = 3.5 \text{ A/mm}^2$ and a slot fill factor $k_{fill} = 0.4$, and noticing that there are two coils per slot, the slot area A_{slot} is as follows:

$$A_{slot} = \frac{2 \cdot n_c \cdot I}{j_{cn} \cdot k_{fill}} = \frac{\sqrt{2} \cdot 0.855 \cdot 10^3}{3.5 \cdot 10^6 \cdot 0.4} = 0.863 \cdot 10^{-3} \text{ m}^2 \quad (11.71)$$

The slot detailed geometry may now be calculated if the stator pole (tooth) width is first found. As the current loading in such low-speed machines tends to be large, and the coupling between the PM and the coils is not so strong, an analytical dimensioning of its magnetic core is hardly practical, although apparently standard methods could be used. The rather low rated current density is a safety precaution for the torque capability (above 200 Nm) exploration by FEM.

11.4.2 FEM Analysis of Pole-PM FRM at No Load

The geometry with PM height $h_{PM} = 2.5 \text{ mm}$ and airgap $g = 0.5 \text{ mm}$ was, in fact, obtained after many FEM calculations with thinner and thicker PMs and eventually larger airgaps. The maximum PM flux, calculated in the middle of the coil, was found for $h_{PM} = 2.5 \text{ mm}$ and $g = 0.5 \text{ mm}$ (PM pitch is equal to 10 mm). The PM flux per stator pole with straight rotor poles is given in Figure 11.16. It should be noted that the waveforms are symmetric and almost sinusoidal [4].

However, the cogging torque, although not large (less than 10% rated torque), for general purposes, may be further reduced. Skewing the rotor by various angles leads to cogging torque reduction (Figure 11.17a). For 1.8° mechanical skewing, the best result is obtained (Figure 11.17b).

A 7% loss in the maximum PM flux per phase is encountered with 1.80 mechanical degrees of skewing (Figure 11.17c). We consider this to be acceptable, as the cogging torque was reduced below 1.5%, which qualifies FRM for high performance. The skewing has the additional effect of shifting the phase PM flux (as expected). This effect is important when designing the control system.

After filtering the phase PM fluxes obtained from FEM, their derivation with rotor position was performed. As demonstrated in Figure 11.18a and Figure 11.18b, there are serious grounds to use vector (sinusoidal) control, as the emfs are proportional to $d\Phi/d\theta$, and they are almost sinusoidal.

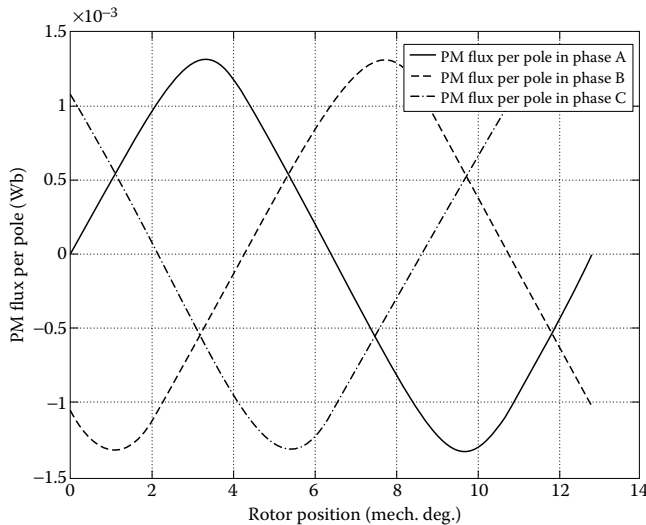


FIGURE 11.16 Permanent magnet (PM) flux per pole vs. rotor position.

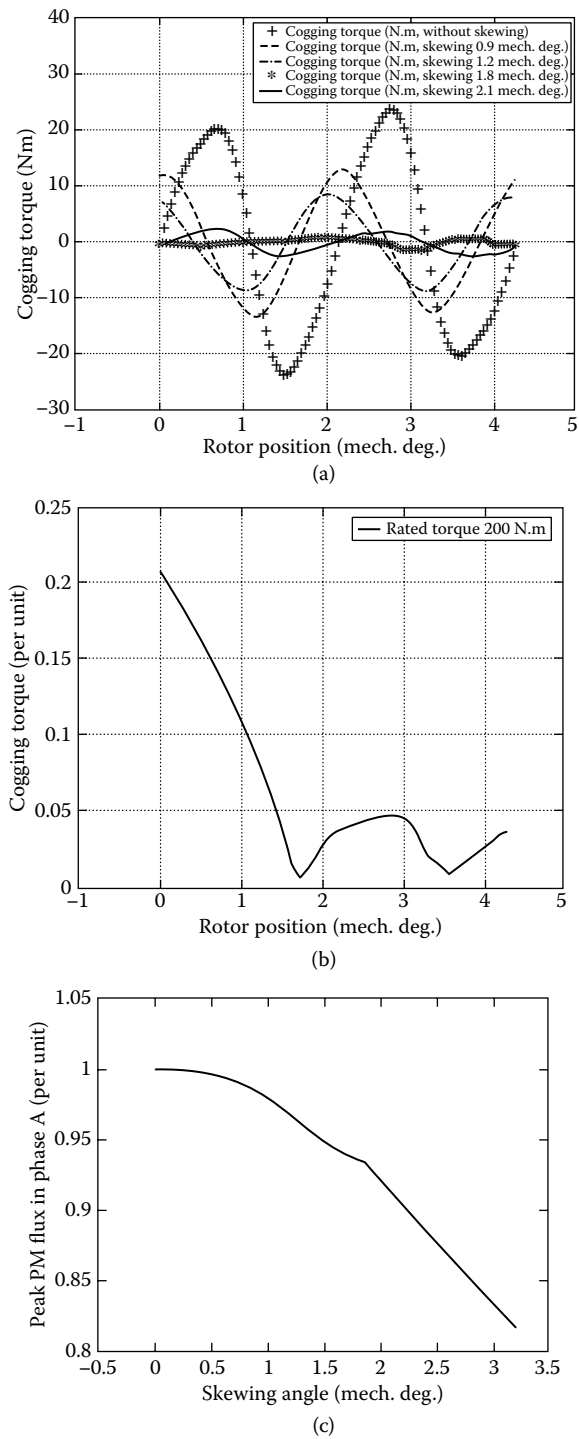


FIGURE 11.17 Cogging torque for different skewing angles: (a) cogging torque vs. position, (b) cogging torque vs. skewing angle, and (c) peak permanent magnet (PM) flux in a phase coil vs. skewing angle.

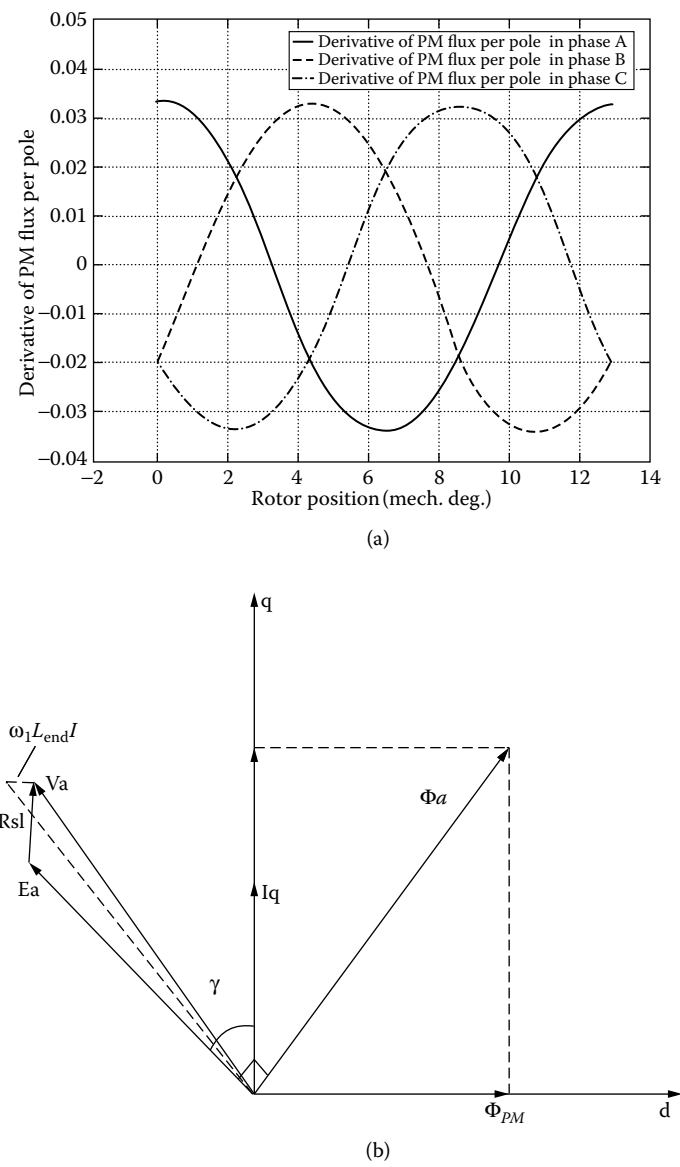


FIGURE 11.18 Waveforms of $d\Phi_{PM}/d\theta_r$ per pole in phases A, B, and C: (a) fundamental and third harmonic and (b) vector diagram with pure I_q control.

11.4.3 FEM Analysis at Steady State on Load

We may simulate steady-state load conditions by FEM if we let sinusoidal currents flow in the three phases. Therefore, in fact, by gearing the currents to rotor position such that the current in phase A is maximum when the PM flux in phase A is zero, we produce pure q axis control. Figure 11.19a shows the phase currents vs. position. They are “in phase” with the flux derivatives (Figure 11.18a).

For various rotor positions, the flux per pole of phases A, B, and C with all currents present with instantaneous values, which are position dependent, FEM flux distribution was calculated. The flux per pole derivatives with θ_r for the three phases are determined and shown in Figure 11.19b. It should be noted that, as expected, the on-load total flux derivatives are not in phase with the currents, as they, in

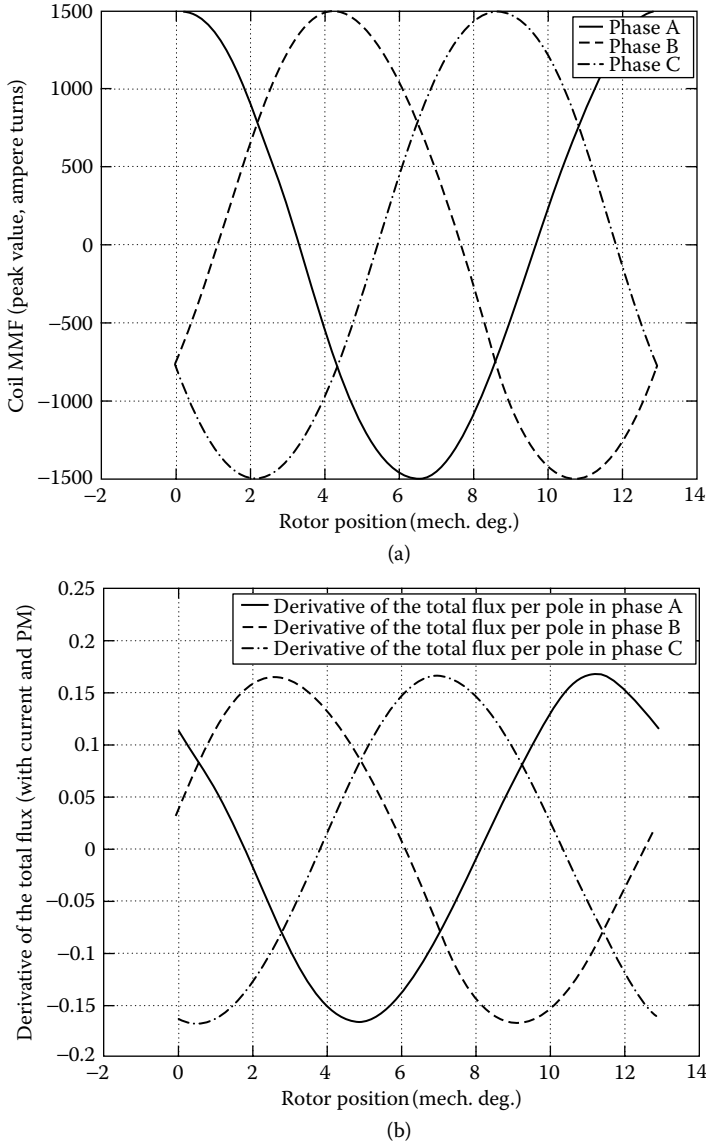


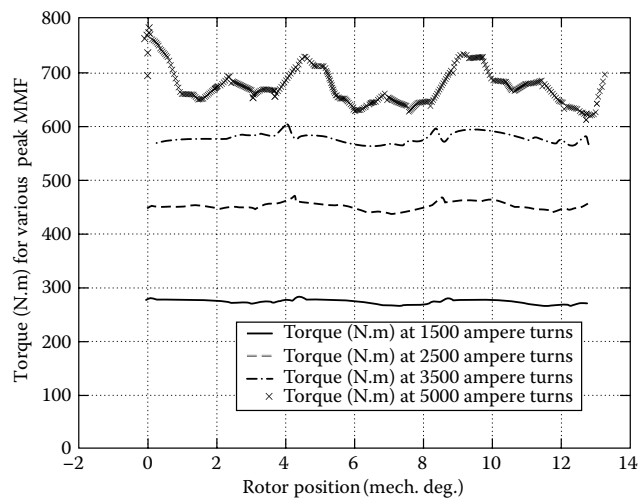
FIGURE 11.19 (a) Phase currents and (b) total flux derivatives with respect to rotor position θ_r .

fact, represent the total derivatives, and because current varies, also. That is to say, the total $E(A, B, C)$ under steady state is proportional to $d\Phi_{A, B, C}/d\theta_r$:

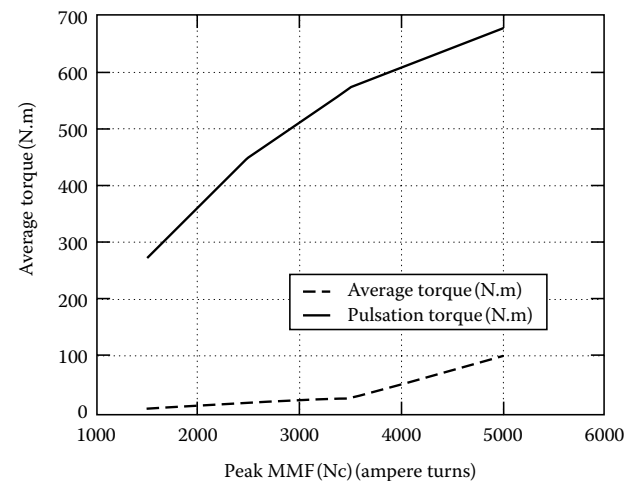
$$E_{A,B,C}^t(\theta_r) = \frac{-d\phi_{A,B,C}(\theta_r)}{d\theta_r} \cdot \frac{N_s}{3} \cdot n_c \cdot 2 \cdot \pi \cdot n \text{ V (rms)} \quad (11.72)$$

Taking the maximum value of $E_A^t(\theta_r)$ and dividing it by $\sqrt{2}$, we may add the voltage phasor RI_A (Figure 11.17c) and get the phase voltage V_A :

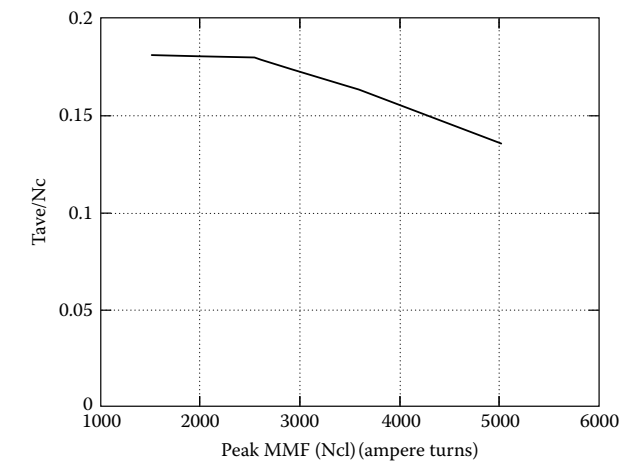
$$V_A \approx \sqrt{\left(E_A^t(\theta_r)_{\max} + R_s \cdot I \cdot \cos \gamma\right)^2 + R_s \cdot I^2 \cdot \sin^2 \gamma} \quad (11.73)$$



(a)



(b)



(c)

FIGURE 11.20 Curves representing current-related torque vs. position for various coil peak magnetomotive force (mmf) (sinusoidal currents in all phases): (a) and (b) average and pulsation torque vs. mmf, (c) average torque/pole mmf, and (d) cogging torque vs. position.

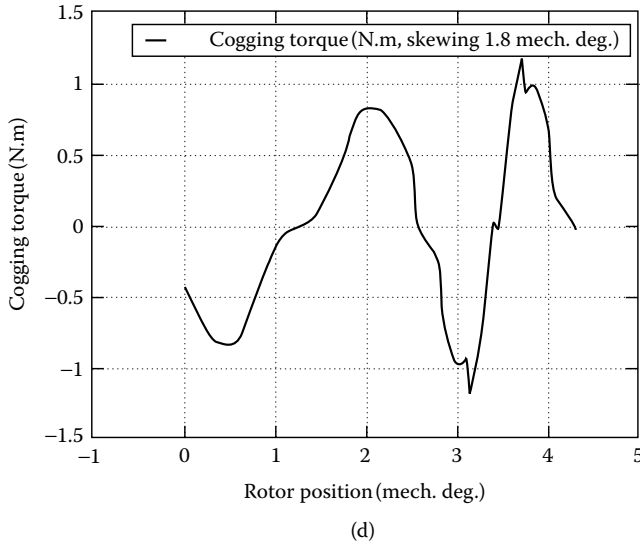


FIGURE 11.20 (Continued).

The only component missing is the end-connection leakage inductance component L_{end} . This inductance produces an additional voltage drop $\omega_l \cdot L_{end} \cdot I$ ($\omega_l = 2 \cdot \pi \cdot n \cdot N_r = 2 \cdot \pi \cdot f_l$, Figure 11.17b). This way, the steady-state characteristics may be calculated for various current levels in the machine. The flux per pole in phase A for a coil peak mmf of 1500, 2500, 3500, and 5000 Atturns demonstrates the effect of saturation on performance. This is more evident in instantaneous torque $T_e(t)$:

$$T_e(t) = \frac{N_s}{3} \cdot n_c \cdot \left[\frac{d\phi_A^i(\theta_r)}{d\theta_r} \cdot i_A(\theta_r) + \frac{d\phi_B^i(\theta_r)}{d\theta_r} \cdot i_B(\theta_r) + \frac{d\phi_C^i(\theta_r)}{d\theta_r} \cdot i_C(\theta_r) \right] \quad (11.74)$$

The results are shown in Figure 11.20a through Figure 11.20d. A few remarks are in order.

- Tangential force densities up to 7.108 N/cm² are practical, but at a low power factor (less than 0.3 in our case).
- The current-related torque pulsations increase with current due to saturation and the presence of small reluctance torque (Figure 11.20a).
- The torque/pole ampere turns (Figure 11.20c) decreases with current, especially for mmfs above 3500 ampere turns/pole.
- Even for 5000 ampere turns/pole, the PMs are not demagnetized (Figure 11.21).
- The current-related torque pulsations are 3% of the average torque at 3500 ampere turns/pole.
- If the cogging torque peak value is considered, the total torque pulsation/amplitude is 3.01% at 3500 ampere turns. This is not a large value and may be further reduced by introducing a small harmonic in the current reference waveform to compensate for it.

As the PM height $h_{PM} = 2.5$ mm, and the airgap $g = 0.5$ mm, the question arises if, for 5000 ampere turns, the PMs do not get demagnetized. Through FEM, the flux density distributions in the PM (half North Pole and half South Pole) lower and upper borders were calculated (Figure 11.21). The PMs do not get demagnetized mainly because of the large fringing effects in a large magnetic gap structure. In other words, the not so good coupling between the PMs and the stator coils allows for a high current loading.

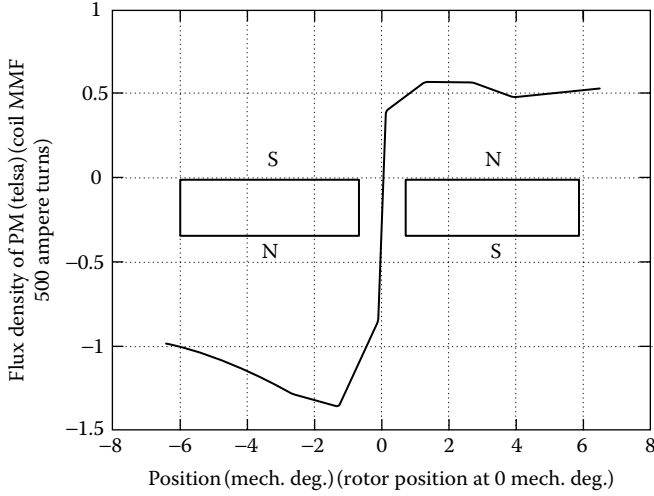


FIGURE 11.21 Flux density on the permanent magnet (PM) upper and lower border vs. position for 5000 A turns/pole and $\theta_r = 0$.

While the torque (tangential force) density is important, so is the ratio between stator losses and torque (W/Nm). For our case, and rated current ($j_{con} = 3.5 \text{ A/mm}^2$), the coil loss is as follows:

$$\begin{aligned}
 P_{coil} &= 3 \cdot R_s \cdot I_n^2 = 3 \cdot \rho_{co} \cdot \frac{l_{coil} \cdot j_{con}}{n_c \cdot I_n} \cdot \frac{N_s}{3} \cdot n_c^2 \cdot I_n^2 = 3 \cdot k_R \cdot n_c^2 \cdot I_n^2 \\
 &= 3 \cdot 2.1 \cdot 10^{-8} \cdot 0.46 \cdot 3.5 \cdot 10^6 \cdot 4 \cdot \frac{855}{\sqrt{2}} = 244.95 \text{ W}
 \end{aligned} \quad (11.75)$$

and

$$l_{coil} \approx 2 \cdot (l_{stack} + 2 \cdot n_p \cdot \tau_{PM}) = 2 \cdot (0.19 + 2 \cdot 2 \cdot 0.1) = 0.46 \text{ m} \quad (11.76)$$

The total torque T_{en} is considered to be 200 Nm, and, thus, the ratio $P_{coil}/T_{en} = 244.95/200 = 1.225 \text{ W/Nm}$. This is a very good result, but the force density is moderate:

$$f_{tn} = \frac{2 \cdot T_{en}}{\pi \cdot D_r^2 \cdot l_{stack}} = \frac{2 \cdot 200}{\pi \cdot 0.179^2 \cdot 0.189^2} = 2.1 \cdot 10^4 \text{ N/m}^2 = 2.1 \text{ N/cm}^2 \quad (11.77)$$

This situation corresponds to a pole peak value of mmf of 855 ampere turns. For 5000 ampere turns/pole, the torque is 677 Nm (7.1085 N/cm²), so the torque increases by 3.385 times, but the losses increase $(5000/855)^2 = 34.2$ times. Also, the current density for peak torque is 16.95 A/mm². Such a current density may be sustained for continuous operation only with liquid stator cooling.

For the peak torque, the efficiency with core and PM eddy current losses neglected, as they are relatively small, is as follows:

$$\eta_k \approx \frac{T_{ek} \cdot \Omega_n}{T_{ek} \cdot \Omega + p_{coil}} = \frac{677 \cdot 2 \cdot \pi \cdot \frac{128}{60}}{677 \cdot 2 \cdot \pi \cdot \frac{128}{60} + 244.95 \cdot 23.47} = 0.52 \quad (11.78)$$

For rated conditions, however,

$$\eta_n \approx \frac{T_{en} \cdot \Omega_n}{T_{en} \cdot \Omega + p_{coil}} = \frac{200 \cdot 2 \cdot \pi \cdot \frac{128}{60}}{200 \cdot 2 \cdot \pi \cdot \frac{128}{60} + 244.95} = 0.91 \quad (11.79)$$

The ideal power factor angle ϕ_1 is (for $n_c I \sqrt{2} = 855$ Aturns, from Equation 11.5),

$$\phi_1 = \tan^{-1} \frac{\omega_1 \cdot L_s \cdot I \cdot \sqrt{2}}{E_{peak}} = \tan^{-1} 2.2 = 65^\circ$$

or $\cos \phi_1 = 0.41$. Considering the low speed of 128 rpm, the efficiencies are reasonable. Note that the 0.52 efficiency at very heavy current load implies a large voltage drop along stator resistance, to be considered when calculating the number of turns/coil (n_c) for a given inverter voltage value. The power factor is already low, as in general, the magnetic airgap is small, and $L_s I_s$ is still large.

11.4.4 FEM Computation of Inductances

As already mentioned, the inductance needed for the circuit model is the transient inductance:

$$L_{AA_t} = \left(\frac{\Delta \phi_A}{\Delta i_A} \right); L_{AB_t} = \left(\frac{\Delta \phi_B}{\Delta i_A} \right); \theta_r, i_B, i_C = \text{const.} \quad (11.80)$$

Therefore, only the current amplitude in phase A was modified by a relatively small value, while the values of currents in the other phases remain unchanged. When the rotor position is varied, the instantaneous values of the currents in all phases vary accordingly. The self-inductance and mutual transient inductance, L_{AA_t} and L_{AB_t} , respectively, are shown in Figure 11.22a and Figure 11.22b for two-phase current amplitudes (corresponding to 2500 and 2600 peak mmf/pole). Two remarks are appropriate [4]:

- The variation of transient inductance with θ_r may be neglected.
- The mutual transient inductance is almost ten times smaller than the self-inductance.

Repeating the FEM computation for a low value of current in the machine, the average self-inductance and mutual inductance are shown to vary notably with current due to magnetic saturation (Figure 11.23). To check the necessity of considering all currents present when calculating the transient inductances, the computation has also been done with current in phase A only. Figure 11.23 shows clearly that a correct estimation of transient inductance requires all phases to be energized. The same is true when calculating the torque from emfs and currents.

For the circuit model, we may use either the phase coordinate or the d - q model. Let us first define the phase coordinate model:

$$i_{A,B,C} \cdot R_s - V_{A,B,C} = E_{A,B,C}(\theta_r, i_s) - L_t(i_s) \cdot \frac{di_{A,B,C}}{dt} \quad (11.81)$$

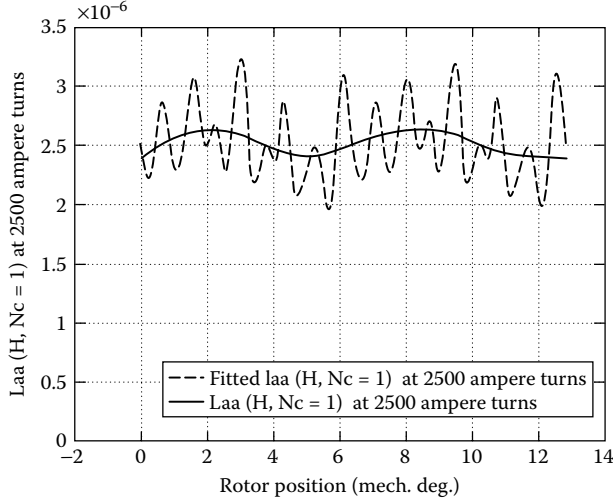
$$E_{A,B,C} = - \left(\frac{d\phi_{A,B,C}(\theta_r, i_s)}{d\theta_r} \right)_{i_s} \cdot \frac{N_s}{3} \cdot n_c \cdot 2 \cdot \pi \cdot n \quad (11.82)$$

11.4.5 Inductances and the Circuit Model of FRM

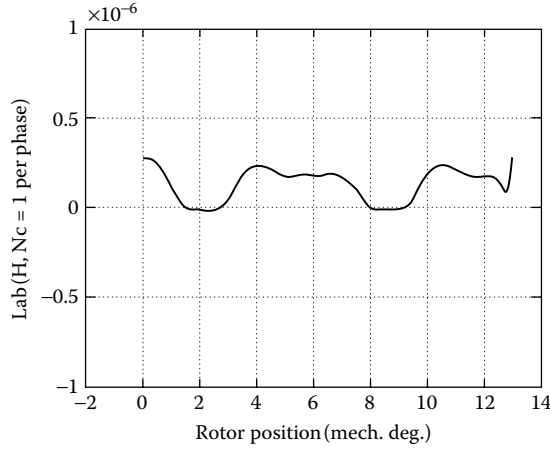
First, the synchronous inductance is

$$L_t(i_s) = L_{end} + [L_{AA_t}(i_s) - L_{AB_t}(i_s)] \quad (11.83)$$

$$i_s = \left| \frac{2}{3} \cdot [i_A(t) + i_B(t) \cdot e^{j(2\pi/3)} + i_C(t) \cdot e^{-j(2\pi/3)}] \cdot e^{-jN_r \theta_r} \right| \quad (11.84)$$



(a)



(b)

FIGURE 11.22 Self-inductance and mutual inductance with all currents in phases A, B, and C considered: (a) self-inductance and (b) mutual inductance.

We use here the current space vector absolute value to account for saturation with the A, B, C model. Also, analytical approximations in total emf and inductance dependence on θ_r and i_s are required:

$$\frac{d\phi(i_s, \theta_r)}{d\theta_r} = (a_0 + a_1 \cdot n_c \cdot I_s - a_2 \cdot n_c^2 \cdot I_s^2) \cdot [\cos(N_r \cdot \theta_{rA,B,C}) + a_3 \cdot \cos(3 \cdot N_r \cdot \theta_{rA,B,C})] \quad (11.85)$$

where

$$\theta_{rA,B,C} = \theta_r + \gamma_0(n_c \cdot I_s) - (k-1) \cdot \frac{2 \cdot \pi}{3}, k=1,2,3$$

and

$$\begin{aligned} \gamma_0(n_c \cdot I_s) &= \gamma_{\theta 0} + c_1 \cdot n_c \cdot I_s + c_2 \cdot n_c^2 \cdot I_s^2 \\ L_t(i_s) &\approx [L_{end} + b_1 - b_2 \cdot n_c^2 \cdot I_s^2] \cdot n_c^2 \end{aligned} \quad (11.86)$$

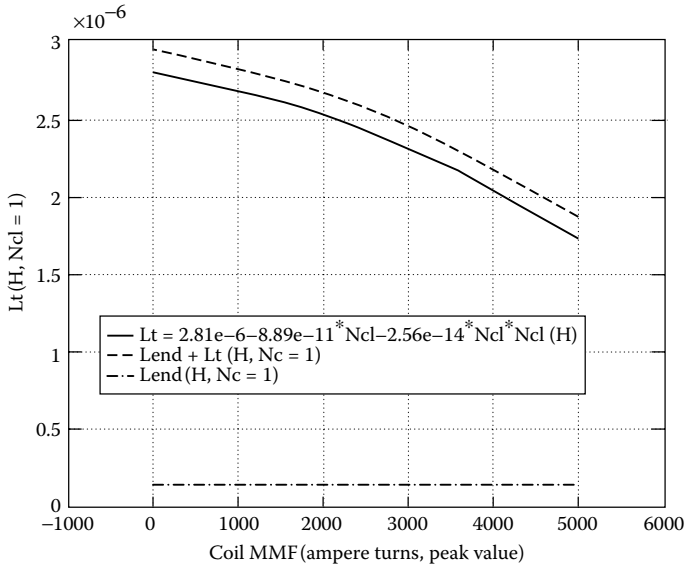


FIGURE 11.23 Average transient inductance dependence on current.

The torque equation T_e is as follows:

$$T_e = (E_A \cdot i_A + E_B \cdot i_B + E_C \cdot i_C) \cdot \frac{1}{2 \cdot \pi \cdot n} \quad (11.87)$$

The motion equations complete the model:

$$J \cdot 2 \cdot \pi \cdot \frac{dn}{dt} = T_e - T_{load}; \quad \frac{d\theta_r}{dt} = n \quad (11.88)$$

11.4.6 The d - q Model of FRM

The study of transients may be performed in A, B, C coordinates, but for control purposes, the d - q model is required. To simplify the d - q model, we will neglect the third harmonic in the emfs, as it is below 5%. When implementing the control for very low torque pulsation, the current i_q may be supplied an additional component to cancel the already small torque pulsations. The d - q model, in stator coordinates, is simply,

$$\bar{i}_s \cdot R_s - \bar{V}_s = -\frac{d\bar{\psi}_s}{dt} - j \cdot \omega_r \cdot \bar{\psi}_s; \quad \frac{d\bar{\psi}_s}{dt} = L_t \cdot \frac{d\bar{i}_s}{dt} \quad (11.89)$$

$$\begin{aligned} \bar{\psi}_s &= \psi_d + j \cdot \psi_q; \quad \bar{i}_s = i_d + j \cdot i_q; \\ \bar{V}_s &= V_d + j \cdot V_q \end{aligned} \quad (11.90)$$

$$i_s = \sqrt{i_d^2 + i_q^2} \quad (11.91)$$

$$P(\theta_r) = \frac{2}{3} \begin{bmatrix} \cos(-\theta_r) & \cos\left(-\theta_r + \frac{2 \cdot \pi}{3}\right) & \cos\left(-\theta_r - \frac{2 \cdot \pi}{3}\right) \\ \sin(-\theta_r) & \sin\left(-\theta_r + \frac{2 \cdot \pi}{3}\right) & \sin\left(-\theta_r - \frac{2 \cdot \pi}{3}\right) \end{bmatrix}$$

$$\begin{bmatrix} \omega_r \cdot \psi_d(i_s) \\ \omega_r \cdot \psi_q(i_s) \end{bmatrix} = P(\theta_r) \cdot \begin{bmatrix} E_A(\theta_r, i_s) \\ E_B(\theta_r, i_s) \\ E_C(\theta_r, i_s) \end{bmatrix} \quad (11.92)$$

$$\begin{aligned} \psi_d(i_s) &= L_s(i_s) \cdot i_d + \psi_{PM} \\ \psi_q(i_s) &= L_s(i_s) \cdot i_q \end{aligned} \quad (11.93)$$

$$T_e = \frac{2}{3} \cdot N_r \cdot (\psi_d \cdot i_q - \psi_q \cdot i_d) \quad (11.94)$$

$$J \cdot 2 \cdot \pi \cdot \frac{dn}{dt} = T_e - T_L; \quad \frac{d\theta_r}{dt} = 2 \cdot \pi \cdot n \cdot p \quad (11.95)$$

Notice that L_s is the average normal steady-state inductance. L_s and the transient inductance L_t both depend on i_s due to saturation (Figure 11.23):

$$\bar{V}_s = \frac{2}{3} \cdot \left(V_A(t) + V_B(t) \cdot e^{j(2\pi/3)} + V_C(t) \cdot e^{-j(2\pi/3)} \right) \quad (11.96)$$

As expected, with sinusoidal $E(\theta_r)$, the d - q model will not exhibit θ_r other than in the Park transformation. The magnetic saturation is considered according to Equation 11.83 through Equation 11.86.

Example 11.3

Consider the preliminary design of a FRM as generator for a torque $T_{en} = 200$ kNm at a speed $n_n = 30$ rpm for a frequency $f_n \approx 100$ Hz.

Solution

Making use of Equation 11.63, with $f_m = 2.66$ N/cm² with $\lambda = 0.3$, we obtain directly the interior diameter, D_{ir} :

$$D_{ir} = \sqrt[3]{\frac{2 \cdot T_{en}}{\pi \cdot f_m \cdot \lambda}} = \sqrt[3]{\frac{2 \cdot 200 \cdot 10^3}{\pi \cdot 2.66 \cdot 10^4 \cdot 0.3}} = 2.5158 \text{ m}$$

The stack length $l_{stack} = \lambda \cdot D_{ir} = 0.3 \cdot 2.5158 = 0.7545$ m. This is the same as in Example 11.1 for the three-phase TFM.

For 100 Hz, the number of rotor salient poles N_r is as follows (Equation 11.56):

$$N_{ri} = \frac{f_{mi}}{n_n} \approx \frac{100}{\frac{30}{60}} = 200$$

We are choosing the same PM material, NdFeB, with $B_r = 1.3$ T and $\mu_{rem} = 1.05$ at 100°C.

The number of stator large poles (coils) N_s and the number of PM poles n_{pp} per stator large pole are as follows:

$$N_s \cdot \left(2 \cdot n_{pp} + \frac{2}{3} \right) = 2 \cdot N_r$$

and thus, with $N_s = 6 \cdot k$ and $n_{pp} = 4$, we obtain $N_s = 48$ for $N_r = 208$. Consequently, the PM pole width τ_{PM} is as follows:

$$\tau_{PM} = \frac{\pi \cdot D_{ir}}{2 \cdot N_r} = \frac{\pi \cdot 2.515}{2 \cdot 208} = 0.01898 \text{ m}$$

and the frequency $f_n = n_n \cdot N_r = 30/60 \cdot 208 = 104 \text{ Hz}$.

As the configuration is more rugged (it resembles the SRM topology), we may adopt a slightly smaller airgap of $g = 3 \text{ mm}$.

We then assign the PM radial height h_{PM} a value, which is about three times the airgap $h_{PM} = 3 \cdot g = 9 \text{ mm}$, to provide for high enough ideal maximum PM flux density in the airgap:

$$B_{gPM\text{Maxi}} = B_r \cdot \frac{h_{PM}}{h_{PM} + g} = 1.3 \cdot \frac{9}{9 + 3} = 0.975 \text{ T}$$

The actual maximum PM flux density in the airgap $B_{gPM\text{Max}}$ is as follows:

$$B_{gPM\text{Max}} = B_{gPM\text{Maxi}} \cdot \frac{1}{(1 + k_{fringe}) \cdot (1 + k_s)} \quad (11.97)$$

with the fringing coefficient $k_{fringe} = 2.33$ and the magnetic core contribution $k_s = 0.1$, $B_{gPM\text{Max}}$ becomes

$$B_{gPM\text{Max}} = 0.975 \cdot \frac{1}{(1 + 2.33) \cdot (1 + 0.1)} = 0.266 \text{ T}$$

Note that this is about the same as for TFM in Example 11.1.

This seems a small value, but at this small pole pitch, it may be justified, given the large gap imposed by the large rotor diameter.

The peak value of the phase emf, with N_s coils in series, and each having n_c turns, is as follows (Equation 11.68):

$$\begin{aligned} E_m &= \frac{N_s}{3} \cdot n_c \cdot 2 \cdot \pi \cdot n \cdot l_{stack} \cdot \frac{\pi \cdot D_r}{2} \cdot n_{pp} \cdot B_{gPM\text{Max}} \\ &= \frac{48}{3} \cdot 2 \cdot \pi \cdot \frac{30}{60} \cdot 0.7545 \cdot \pi \cdot \frac{2.515}{2} \cdot 4 \cdot 0.266 \cdot n_c = 159.2 \cdot n_c \end{aligned}$$

The RMS ampere turns per coil $n_c I_n$ for rated torque are as follows (Equation 11.70):

$$n_c \cdot I_n = \frac{2}{3} \cdot \frac{n_c}{\sqrt{2}} \cdot T_{en} \cdot \frac{2 \cdot \pi \cdot n}{E_m} = \frac{2 \cdot 200 \cdot 10^3 \cdot 2 \cdot \pi \cdot 1/2}{3 \cdot \sqrt{2} \cdot 159.2} = 1865 \text{ Aturns(rms)/coil} \quad (11.98)$$

The area required in the slot to accommodate two coils A_{slot} is

$$A_{slot} = \frac{2 \cdot n_c \cdot I_n}{j_{con} \cdot k_{fill}} = \frac{2 \cdot 1865}{9 \cdot 10^6 \cdot 0.4} = 1.036 \cdot 10^{-3} \text{ m}^2 = 1036 \text{ mm}^2$$

where $j_{con} = 9 \text{ A/mm}^2$ and $k_{fill} = 0.4$.

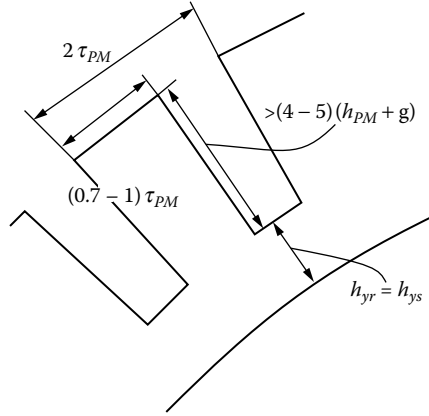


FIGURE 11.25 Rotor salient-pole geometry.

The copper losses for rated current (torque) are as follows:

$$p_{con} = 3 \cdot R_s \cdot I_n^2 = 3 \cdot 3.14 \cdot 10^{-3} \cdot (n_c \cdot I_n)^2 = 9.42 \cdot 10^{-3} \cdot 1865^2 = 32.76 \text{ kW}$$

Note that the simplicity of the manufacturing process in FRM is paid for dearly in copper losses, which increase from 13.70 kW in the TFM to 32.76 kW in the FRM.

As there are no aluminum carriers to hold the stator core, their eddy current losses are also absent in FRM, but still the copper losses of FRM are larger in a machine with the same size and torque. The machine inductance L_s is again made up of two components: the airgap inductance L_m and the leakage inductance L_{sl} . The leakage inductance is moderate, as the slot aspect ratio h_{sl}/b_1 is small, and thus, we concentrate on the airgap inductance:

$$L_m \approx \frac{N_s}{3} \cdot \mu_0 \cdot n_c^2 \cdot \frac{\tau_{PM} \cdot n_{pp}}{g + h_{PM}} \cdot \frac{1 + k_f}{1 + k_s} \cdot l_{stack} \quad (11.99)$$

k_f is a fringing effect coefficient $k_f < 0.2$, which accounts for the large airgap above the rotor interpole contribution to the coil self-flux linkage:

$$L_m = \frac{48}{3} \cdot 1.256 \cdot 10^{-6} \cdot \frac{0.01898 \cdot 4}{(3+9) \cdot 10^{-3}} \cdot \frac{1+0.2}{1+0.1} \cdot 0.7545 \cdot n_c^2 = 0.1046 \cdot 10^{-3} \cdot n_c^2$$

Considering that the leakage inductance represents 10% of L_m , $L_s = (1 + 0.1) \cdot 0.115 \cdot 10^3 \cdot n_c^2$. The power factor angle for pure I_q control is, again (Equation 11.5),

$$\varphi_1 = \tan^{-1} \left(\frac{\omega_1 \cdot L_s \cdot I \cdot \sqrt{2}}{E_m} \right) = \tan^{-1} \left(\frac{2 \cdot \pi \cdot 104 \cdot 0.115 \cdot 10^{-3} \cdot n_c^2 \cdot I \cdot \sqrt{2}}{159.6 \cdot n_c} \right) = \tan^{-1} 1.2375 = 51^\circ$$

$$\cos \varphi_1 = 0.6285$$

Remember that the power factor for the same output conditions was 0.933 for the TFM. We also have to point out that the PM flux fringing ratio of 0.33 (that is, 33% of ideal PM flux reaches the coils) is a bit too severe for the FRM, because the rotor poles are one pole pitch apart tangentially, while for the TFM, the axial distance between neighboring U- and I-shape cores is

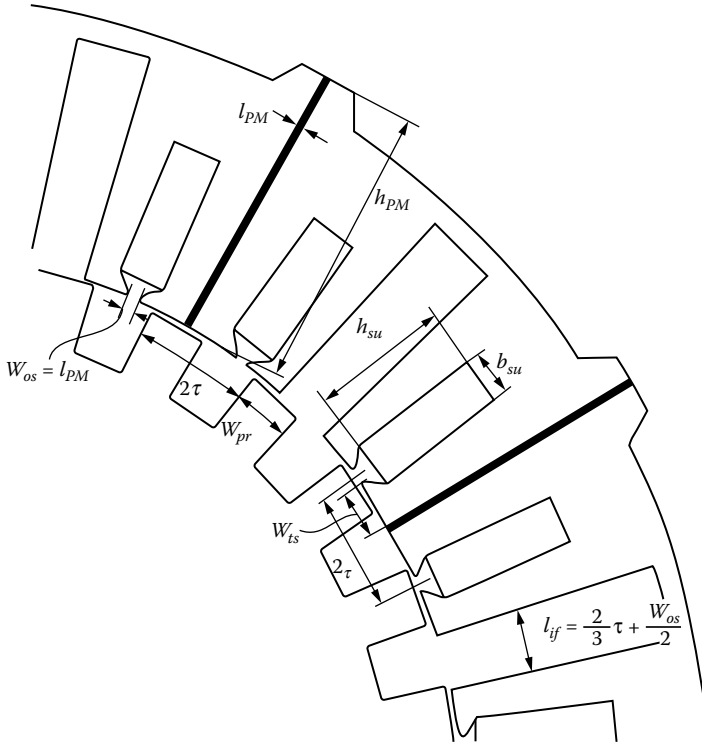


FIGURE 11.26 Flux reversal generator (FRG) with stator permanent magnet (PM) flux concentration.

two to three times smaller. Finally, the PM and airgap height were reduced by 25% for FRM. In-depth FEM studies are required to document the best solution of the two. However, with four airgaps per coil, TFM is expected to show smaller inductance.

Example 11.4: FRM with Stator PM-Flux Concentration

Let us consider again a direct-driven wind generator system operating again at $T_{en} = 200$ kNm and $n = 30$ rpm, $f_n \approx 100$ Hz. The design of the magnetic circuit, losses, and power factor angle for pure I_q control are required.

Solution

The machine geometry (Figure 11.6) is represented with one pole only (Figure 11.26). Making use of the PM flux concentration, we adopted a larger force density f_{tn} than in previous design examples to offset the effect of the inductance increase due to the fringing effect: $f_{tn} = 6$ N/cm².

To yield the same diameter as before, $\lambda = l_{stack}/D_{is}$ is decreased accordingly to $\lambda = 0.133$:

$$D_{is} = \sqrt[3]{\frac{2 \cdot T_{en}}{\pi \cdot f_{tn} \cdot \lambda}} = \sqrt[3]{\frac{2 \cdot 200 \cdot 10^3}{\pi \cdot 6 \cdot 10^4 \cdot 0.133}} = 2.515 \text{ m}$$

The stack length is, however,

$$l_{stack} = \lambda \cdot D_{is} = 0.133 \cdot 2.515 = 0.3345 \text{ m}$$

The number of rotor salient poles for around 100 Hz should be around $N_r = 200$, calculated as follows (Figure 11.26):

$$2 \cdot N_r \cdot \tau = 6 \cdot k \cdot \left(4 \cdot \tau + \frac{2 \cdot \tau}{3} \right) = 28 \cdot k \cdot \tau \quad (11.100)$$

with $N_r = 186$; $k = 14$; $f_1 = N_r \cdot n = 186 \cdot 30/60 = 93$ Hz.

The airgap $g = 2.0$ mm, and the pole pitch τ is

$$\tau = \frac{\pi \cdot D_{is}}{2 \cdot N_r} = \frac{\pi \cdot 2.515}{2 \cdot 186} = 0.02014 \text{ m}$$

The slot opening, W_{os} , equal to the PM thickness, l_{PM} , is considered to be 8 mm. So, the stator tooth $W_{ts} = \tau - W_{os} = 20 - 8 = 12$ mm. The rotor salient pole W_{pr} may be as wide as the pole pitch or smaller:

$$W_{pr} = (0.8 \div 1.0) \cdot \tau \quad (11.101)$$

The ideal PM flux density in the airgap B_{gPMMax} is now as follows:

$$B_{gPMMax} = \frac{B_r}{\frac{W_{ts}}{h_{PM}} + \frac{\mu_{rec}}{\mu_0} \cdot \frac{2 \cdot g}{\tau_{PM}}} \quad (11.102)$$

This expression is derived from

$$\begin{aligned} B_m \cdot h_{PM} &= B_g \cdot W_{ts} \\ H_m \cdot l_{PM} + \frac{B_g}{\mu_0} \cdot 2 \cdot g &= 0 \\ B_m &= B_r + \mu_{rec} \cdot H_m \end{aligned} \quad (11.103)$$

with $B_r = 1.3$ T and $\mu_{rec} = 1.05 \cdot \mu_0$.

The fringing effect will reduce the ideal PM airgap flux density to a smaller value, as in previous cases:

$$B_{gPMMax} = \frac{B_{gPMMaxi}}{(1 + k_{fringe}) \cdot (1 + k_s)} \quad (11.104)$$

The fringing coefficient depends again on the airgap g/W_{ts} ratio, l_{PM}/g ratio, and on the degree of saturation. With $k_{fringe} = 1.5$, $k_s = 0.2$, we set the actual airgap flux density B_{gPMMax} at a reasonable value, say $B_{gPMMax} = 0.7$ T.

In this case, from Equation 11.104,

$$B_{gPMMaxi} = 0.7 \cdot (1 + 1.5) \cdot (1 + 0.2) = 2.1 \text{ T}$$

This is a high value, but remember that it is only a theoretical one. From Equation 11.102, we may size the PM height h_{PM} :

$$2.1 = \frac{1.3}{\frac{12}{h_{PM}} + \frac{1.05 \cdot \mu_0}{\mu_0} \cdot \frac{2 \cdot 20}{8}}; h_{PM} = 120 \text{ mm}$$

The PM maximum flux per coil turn, Φ_{PMMaxc} , is as follows:

$$\phi_{PMMaxc} = B_{gPMMax} \cdot W_{ts} \cdot l_{stack} = 0.7 \cdot 0.012 \cdot 0.3345 = 2.81 \cdot 10^{-3} \text{ Wb} \quad (11.105)$$

The maximum flux linkage $\psi_{PMphase}$ is

$$\psi_{PMphase} = 2 \cdot k \cdot \phi_{PMMaxc} \cdot n_c = 2 \cdot 14 \cdot 2.81 \cdot 10^{-3} \cdot n_c = 0.07867 \cdot n_c \quad (11.106)$$

The emf per phase (peak value) E_m is

$$E_m = 2 \cdot \pi \cdot f_1 \cdot \psi_{PMphase} = 2 \cdot \pi \cdot 93 \cdot 0.07867 \cdot n_c = 45.949 \cdot n_c \quad (11.107)$$

The ampere turns per slot $n_c I_n$ may be calculated from Equation 11.98:

$$n_c \cdot I_n = \frac{2 \cdot n_c}{3 \cdot \sqrt{2}} \cdot T_{en} \cdot \frac{2 \cdot \pi \cdot n}{E_m} = \frac{2 \cdot n_c}{3 \cdot \sqrt{2}} \cdot \frac{200 \cdot 10^3 \cdot 2 \cdot \pi \cdot 30/60}{45.949 \cdot n_c} = 6.4589 \cdot 10^3 \text{ At/turns/coil}$$

The slot width is about equal to the pole pitch (Figure 11.26). $W_{su} = (1 \text{ to } 1.2) \cdot \tau_{PM} \approx 1.1 \cdot 20 = 22$ mm. This way, the stator tooth average width is around half the pole pitch τ_{PM} .

As the total slot height $h_{su} \approx h_{PM} - 2/3 \cdot W_{ts} - 0.005 = 103 \cdot 10^{-3} - 2/3 \cdot 12 \cdot 10^{-3} - 5 \cdot 10^{-3} = 90 \cdot 10^{-3}$ m, the current density required to host the coil j_{con} is as follows:

$$j_{con} = \frac{n_c \cdot I_n}{h_{su} \cdot W_{su} \cdot k_{fill}} = \frac{6458.9}{90 \cdot 22 \cdot 0.5} = 6.524 \text{ A/mm}^2$$

The stator coil resistance is

$$R_{sc} = \rho_{co} \cdot \frac{l_{coil}}{\frac{n_c \cdot I_n}{j_{con}}} \cdot n_c^2 = \frac{2.3 \cdot 10^{-8} \cdot 0.818}{\frac{6458.9}{6.524 \cdot 10^6}} \cdot n_c^2 = 1.9 \cdot 10^{-5} \cdot n_c^2 \quad (11.108)$$

$$l_{coil} = 2 \cdot l_{stack} + 4 \cdot \tau_{PM} + \pi \cdot W_{su} = 2 \cdot 0.3345 + 4 \cdot 0.020 + \pi \cdot 0.022 = 0.818 \text{ m}$$

For the entire phase,

$$R_s = 2 \cdot k \cdot R_{sc} = 2 \cdot 14 \cdot 1.9 \cdot 10^{-5} \cdot n_c^2 = 0.532 \cdot 10^{-3} \cdot n_c^2$$

The stator copper losses are then

$$P_{con} = 3 \cdot R_s \cdot I_n^2 = 3 \cdot 0.532 \cdot 10^{-3} \cdot 6458.9^2 = 66.58 \text{ kW}$$

A few remarks are in order:

- The FRM with stator PM flux concentration cannot appropriately take full advantage of the principle of PM flux magnification.
- The machine size was reduced (the stack length was 0.7545 m for the surface PM stator FRM). This reduction in size is paid for by larger copper losses (66.58 kW for an input power of 628 kW). There is notably more copper in this machine.
- It is possible to redo the design for a smaller force density, but for larger stack length, the same interior stator diameter D_{is} is needed to notably reduce the copper losses. Still, the inductance seems to be higher as each coil “entertains” two airgaps.
- The FRM with stator PM flux concentration seems to be restricted to a smaller number of poles per rotor and small torque machines, where the fabrication costs may be reduced due to its relatively more rugged topology.

11.4.7 Notes on Flux Reversal Generator (FRG) Control

With its easy rotor skewing, an FRM can produce a rather sinusoidal emf and is thus eligible for field orientation or direct power (torque) control as any PM synchronous generator (see [Chapter 10](#) for details and results).

Example 11.5

Let us now consider the FRG with rotor PM flux concentration (Figure 11.27) illustrated here with only a few poles (Figure 11.27). For the same data as in Example 11.3, prepare an adequate design.

Solution

The two stator cores and the rotor core are all made up of standard stamped laminations. Only the external rotor is provided with windings to save room in the rotor ([Figure 11.7](#)) and thus make it more rugged mechanically. Each stator large pole now contains $2 \cdot n_p + 1$ poles and $2 \cdot n_p$ interpoles. It is clearly visible that for maximum flux per large pole, all the PMs in the rotor are active. This is a better PM utilization in addition to PM flux concentration.

First, let us keep $f_m = 6 \text{ N/cm}^2$ and $\lambda = 0.133$. Then, D_{is} (stator interior diameter) stays the same as in Example 11.3; $D_{is} = 2.515 \text{ m}$.

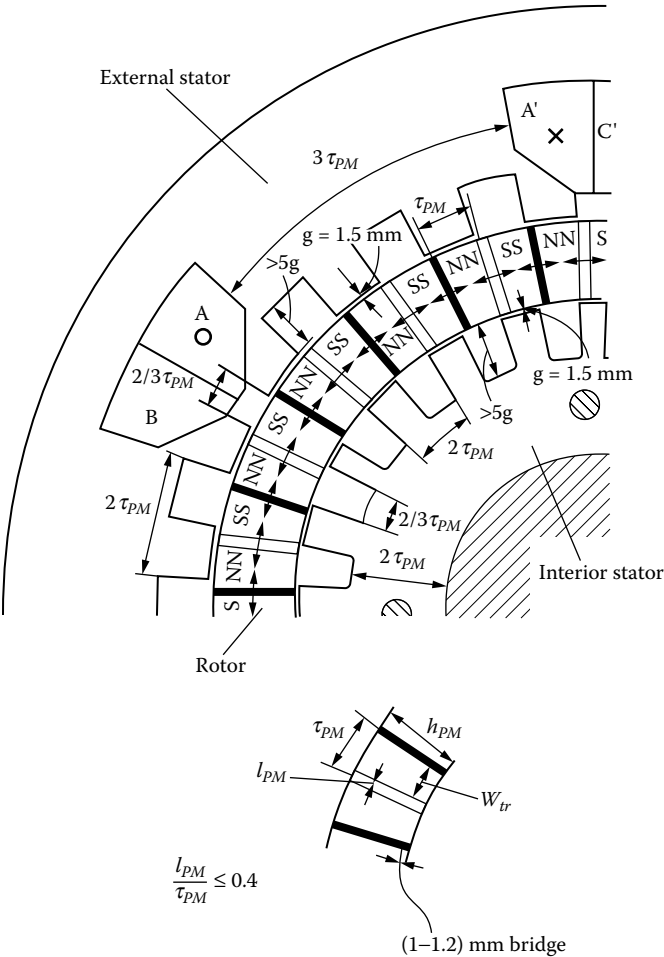


FIGURE 11.27 Flux reversal machine (FRM) with rotor permanent magnet (PM) flux concentration and dual stator.

The number of stator poles $6 \cdot k$ is now as follows:

$$6 \cdot k \cdot \left(2 \cdot n_{pp} + 1 + 2 \cdot n_{pp} + \frac{2}{3} \right) = 2 \cdot N_r \approx 400 \quad (11.109)$$

with $n_{pp} = 2$, $k = 7$, and $2 \cdot N_r = 406$ poles.

So, the PM pole pitch $\tau_{PM} = \pi \cdot D_{is} / 2 \cdot N_r = \pi \cdot 2.515 / 406 = 0.01945$ m.

The ideal PM flux density in the airgap $B_{gPMMaxi}$ is as follows:

$$B_{gPMMaxi} = \frac{B_r}{\frac{W_{tr}}{2 \cdot h_{PM}} + \frac{\mu_{rec}}{\mu_0} \cdot \frac{2 \cdot g}{l_{PM}}}; B_r = 1.3 \text{ T}; \mu_{rec} = 1.05 \cdot \mu_0 \quad (11.110)$$

In contrast to Equation 11.102, the factor $2 \cdot h_{PM}$ instead of h_{PM} is used in Equation 11.110, because two PM magnets cooperate in the rotor tooth W_{tr} .

With the actual airgap PM flux density $B_{gPMMax} = 0.9$ T, the ideal airgap PM flux density $B_{gPMMaxi}$ is as follows:

$$B_{gPMMaxi} = B_{gPMMax} \cdot (1 + k_{fringe}) \cdot (1 + k_s) = 0.9 \cdot (1 + 1.5) \cdot (1 + 0.2) = 2.7 \text{ T} \quad (11.111)$$

With this value in Equation 11.108, the ratio h_{PM}/W_{tr} is obtained (the PM thickness $l_{PM} = 0.4 \cdot \tau_{PM} = 0.4 \cdot 20 = 8$ mm):

$$2.7 = \frac{1.3}{\frac{W_{tr}}{2 \cdot h_{PM}} + \frac{1.05 \cdot \mu_0}{\mu_0} \cdot \frac{3}{8}} \quad (11.112)$$

It follows that $2 \cdot h_{PM}/W_{tr} = 11.4$.

Consequently, the PM radial height $h_{PM} = 11.4 \cdot W_{tr} / 2 = 5.7 \cdot 12 = 68.40$ mm.

This is a notable reduction in PM weight for a better effect in comparison with Example 13.3. The maximum PM flux linkage per phase is obtained by adapting Equation 11.105 and Equation 11.106 for the case in point:

$$\begin{aligned} \Psi_{PMphase} &= 2 \cdot k \cdot W_{ts} \cdot B_{gPMMax} \cdot l_{stack} \cdot (2 \cdot n_{pp} + 1) \cdot n_c \\ &= 2 \cdot 7 \cdot 0.012 \cdot 0.9 \cdot 0.3345 \cdot (2 \cdot 2 + 1) \cdot n_c = 0.2528 \cdot n_c \end{aligned} \quad (11.113)$$

Consequently, the peak value of emf per phase E_m is now calculated from Equation 11.106:

$$E_m = 2 \cdot \pi \cdot f_1 \cdot \Phi_{PMphase} = 2 \cdot \pi \cdot \frac{30}{60} \cdot 203 \cdot 0.2528 \cdot n_c = 161.19 \cdot n_c \quad (11.114)$$

By now, we have all the data necessary to calculate the rated mmf per coil $n_c I_n$ (Equation 11.98):

$$n_c \cdot I_n = \frac{2}{3 \cdot \sqrt{2}} \cdot n_c \cdot T_{en} \cdot \frac{2 \cdot \pi \cdot n}{E_m} = \frac{2}{3 \cdot \sqrt{2}} \cdot n_c \cdot \frac{200 \cdot 10^3 \cdot 2 \cdot \pi \cdot 30/60}{161.19 \cdot n_c} = 1842.1 \text{ Atturns}$$

A current density $j_{con} = 6$ A/mm² may be adopted, as the slot useful area A_{slot} is

$$A_{slot} = \frac{2 \cdot n_c \cdot I_n}{j_{con} \cdot k_{fill}} = \frac{2 \cdot 1842.1}{6 \cdot 10^6 \cdot 0.4} = 1.535 \cdot 10^{-3} \text{ m}^2 = 1535 \text{ mm}^2$$

There is plenty of room to locate such a slot with low height and, thus, with low slot leakage inductance contribution.

The phase resistance R_s (Equation 11.108) is as follows:

$$R_s = 2 \cdot k \cdot \rho_{co} \cdot \frac{l_{coil}}{\frac{n_c \cdot I_n}{j_{con}}} \cdot n_c^2 = \frac{2 \cdot 7 \cdot 2.3 \cdot 10^{-8} \cdot 1.1045}{\frac{1842.1}{6 \cdot 10^6}} \cdot n_c^2 = 1.1684 \cdot 10^{-3} \cdot n_c^2$$

$$l_{coil} \approx 2 \cdot l_{stack} + 2 \cdot \frac{\pi \cdot D_{is}}{6 \cdot k} = 2 \cdot 0.3345 + 2 \cdot \pi \cdot \frac{2.515}{6 \cdot 7} = 1.045 \text{ m}$$

The copper losses p_{con} are

$$p_{con} = 3 \cdot R_s \cdot I_n^2 = 3 \cdot 1.1684 \cdot 10^{-3} \cdot 1842.1^2 = 11.7925 \text{ kW}$$

This displays less copper losses than for the TFM (13.7 kW), while the machine axial total length is about half in the TFM for the same diameter. There is still one more problem: the power factor. For maximum I_q current control, the PM flux is zero in that phase; thus, the airgap inductance is rather large, as the PMs “do not stay in the way” of coil mmf flux.

Consequently,

$$L_m \approx \mu_0 \cdot 2 \cdot k \cdot n_c^2 \cdot (2 \cdot n_{pp} + 1) \cdot \frac{W_{tr} \cdot l_{stack}}{2 \cdot g \cdot (1 + k_s)}$$

$$= 1.256 \cdot 10^{-6} \cdot 2 \cdot 7 \cdot n_c^2 \cdot (2 \cdot 2 + 1) \cdot \frac{0.012 \cdot 3345}{2 \cdot 1.5 \cdot 10^{-3} \cdot (1 + 0.2)} = 0.98 \cdot 10^{-4} \cdot n_c^2 \quad (11.115)$$

$$L_s = (1 + 0.1) \cdot L_m = 1.078 \cdot 10^{-4} \cdot n_c^2$$

Finally, the power factor angle φ_1 is as follows (with $n_c I_n = 1842.1$ Atturns):

$$\varphi_1 = \tan^{-1} \left(\frac{\omega_1 \cdot L_s \cdot I_n \cdot \sqrt{2}}{E_m} \right) = \tan^{-1} \left(2 \cdot \pi \cdot \frac{203}{2} \cdot \frac{1.078 \cdot 10^{-4} \cdot n_c^2 \cdot I_n \cdot \sqrt{2}}{161.2 \cdot n_c} \right) = \tan^{-1} 1.107 = 48^\circ$$

$$\cos \varphi_1 = 0.67$$

Let us try to reduce copper losses further by reducing the number of stator coils with $k = 2$ (four coils per phase). We obtain $n_p = 8$ and $2 \cdot N'_r = 6 \cdot k \cdot (4 \cdot n_p + 1 + 2/3) = 404$. The PM pole pitch τ_{PM} remains about the same at 0.01947 m.

Repeating the design, the PM flux per phase (Equation 11.113) is

$$\psi_{PMphase} = 0.2528 \cdot n_c \cdot \frac{(2 \cdot n'_{pp} + 1) \cdot k'}{(2 \cdot n_{pp} + 1) \cdot k} = 0.2528 \cdot n_c \cdot \frac{(2 \cdot 8 + 1) \cdot 2}{(2 \cdot 2 + 1) \cdot 7} = 0.2455 \cdot n_c$$

and the emf (peak value) E'_m (Equation 11.114) is as follows:

$$E'_m = 2 \cdot \pi \cdot N_r \cdot n \cdot \phi_{PMphase} = 2 \cdot \pi \cdot \frac{202}{2} \cdot 0.2455 \cdot n_c = 155.76 \cdot n_c$$

So, the coil mmf $(n_c I_n)'$ is

$$(n_c \cdot I_n)' = 1842.1 \cdot \frac{161.19}{155.76} = 1906.26 \text{ Atturns(rms)}$$

With the same $j_{con} = 6 \text{ A/mm}^2$, the stator resistance R_s' becomes

$$R_s' = 2 \cdot k \cdot \rho_{co} \cdot \frac{l'_{coil} \cdot n_c^2}{(n_c \cdot I_n)^2} = \frac{2 \cdot 2 \cdot 2.3 \cdot 10^{-8} \cdot n_c^2 \cdot 1.985}{\frac{1906.26}{6 \cdot 10^6}} = 0.575 \cdot 10^{-3} \cdot n_c^2$$

$$l'_{coil} \approx 2 \cdot l_{stack} + \frac{2 \cdot \pi \cdot D_{is}}{6 \cdot k} = 2 \cdot 0.3345 + \frac{2 \cdot \pi \cdot 2.515}{12} = 1.985 \text{ m}$$

The copper losses p'_{con} are now

$$p'_{con} = 3 \cdot R_s' \cdot (n_c \cdot I_n)^2 = 3 \cdot 0.575 \cdot 10^{-3} \cdot 1906.26^2 = 6.266 \text{ kW (about 1\%)}$$

Note that the copper losses were reduced, by half, at the price of thicker stator and iron yokes. Unfortunately, the inductance stays about the same, so the power factor stays about the same, around 0.67.

The following are some final remarks:

- In comparison with the FRM with stator surface PMs, the machine axial length was reduced by half, at the same interior and outer diameter, smaller copper losses, but same power factor.
- The PM weight is only slightly larger than in the TFM with rotor PM flux concentration.
- In comparison with the TFM, the machine length is half at about the same interior and outer stator diameter, the copper losses are only 30% less, but the power factor is 0.67 in comparison with the 0.933 of TFM.
- Reducing the number of stator coils could be another way to further improve FRM performance.
- The FRM is considered more manufacturable than the TFM.
- It could be argued that the double-sided TFM with flux concentration can produce even better results. This is true, but in a less manufacturable topology.
- Though in the numerical examples of this chapter the fringing flux coefficients were chosen conservatively low (0.3 to 0.4), full FEM studies are still required to validate the performance claims to precision, on a case-by-case basis.

11.5 Summary

- This chapter investigates two special PM brushless generators with large numbers of poles and nonoverlapping stator coils.
- One is called the TFM, and it uses circular single coils per phase. The PMs are on the rotor or on the stator, with or without PM flux concentration. The flux paths are generally three dimensional.
- The other one is called the FRM, and it makes use of the switched reluctance machine standard laminated core. The PMs may be placed on top of stator poles in a large even-numbered $2 \cdot n_p$ with alternate polarity.
- In both, TFM and FRM, the PM flux linkage in the stator coils reverses polarity, but, generally, for the same diameter and pole pitch (number of poles per periphery), the TFM coil embraces more alternate PM poles and is destined for better torque magnification.
- As the PM pole pitch decreases with a large number of poles, the PM fringing flux increases to the point that the latter becomes overwhelming. In essence, $\tau_{PM} > g + h_{PM}$, where g is the mechanical gap, and h_{PM} is the PM radial thickness to secure a less than 66% reduction of flux in coil due to fringing. Reductions of only 35 to 40% are reported in Reference 2 for a small machine.
- The large fringing translates into poor usage of PM and core material and high current loading for good torque density, higher copper losses, and lower power factor.
- The torque in Newton meters per watts of copper losses and the power factor $\cos \varphi_1$ are defined as performance indexes, which are independent of speed (frequency), to characterize TFM and FRM. Also, the total cost has to be considered, as these machines use less copper but more iron and PM materials than standard machines.

- The chapter develops preliminary electromagnetic models for TFM and FRM and then uses the same specifications: a 200 kNm, 30 rpm, 100 Hz generator in four different designs — a TFM with surface PMs and three FRM designs (one with surface PM stator, one with stator PM flux concentration, and one with rotor PM flux concentration).
- All four topologies could be designed for the specifications, but the TFM, for the same volume, was slightly better than FRM with the surface PM stator in power factor. However, a reduction in volume by one half, for about the same copper losses and lower power factor, was obtained with the PM rotor flux concentration FRM. It may be argued that the TFM could also be built with PM flux concentration. This is true, but the already low manufacturability of TFM is further reduced this way.
- It is too early to discriminate between TFM and FRM, as one seems slightly better in torque/copper losses for given volume, but the other is notably more manufacturable, for a very high number of poles.
- In terms of control, the FRM is easily capable (through skewing) of controlling sinusoidal emfs and is thus directly eligible for standard field orientation or direct power (torque) control (see [Chapter 10](#) on this issue).
- New PM generator/motor configurations that depart from the standard PM synchronous generators/motors (Chapter 10) are still being proposed in the search for better very low speed direct-driven generator systems with full power electronics control.

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